Electronic technology

principal contents

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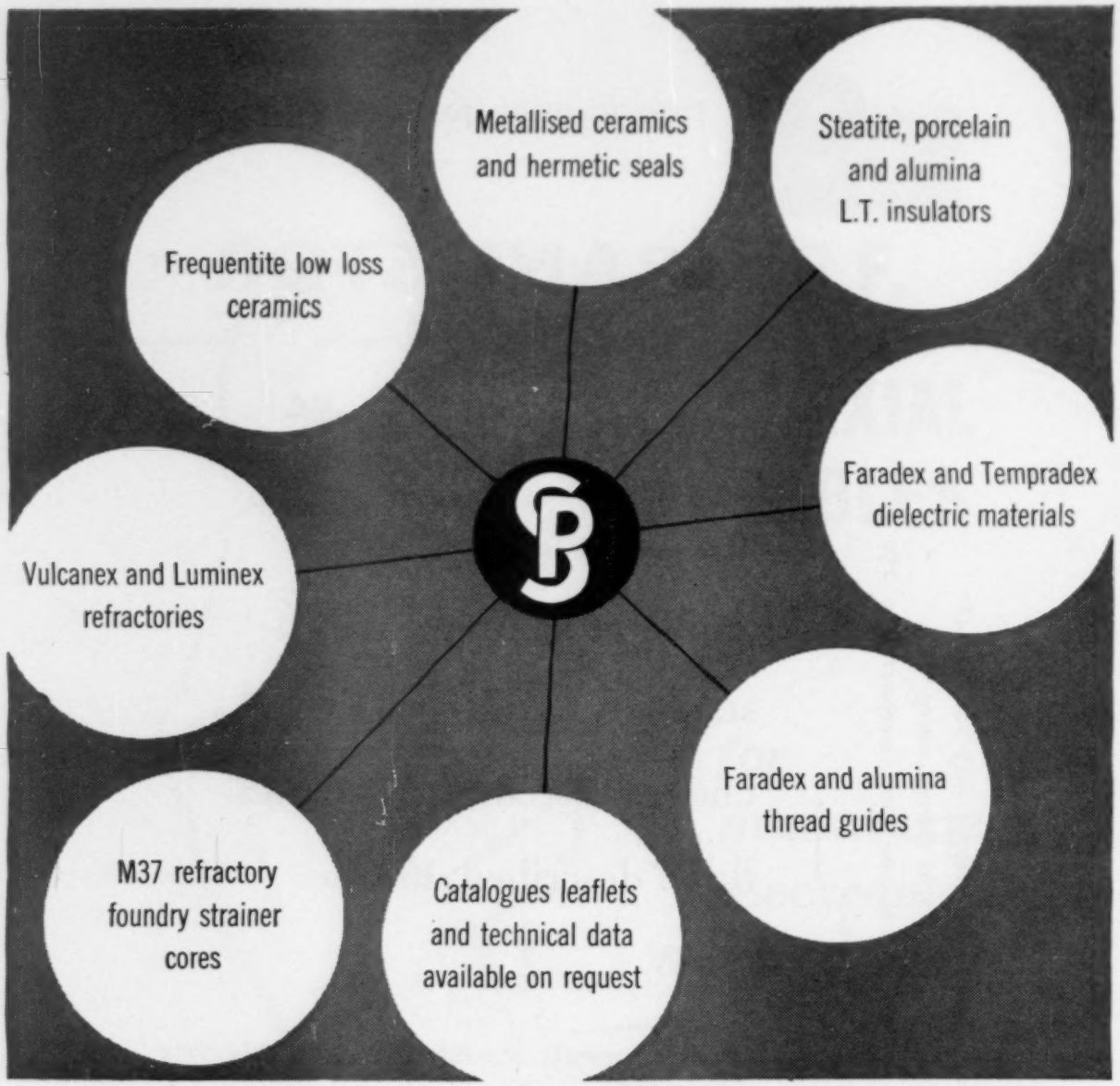
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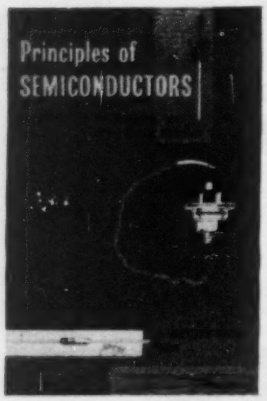

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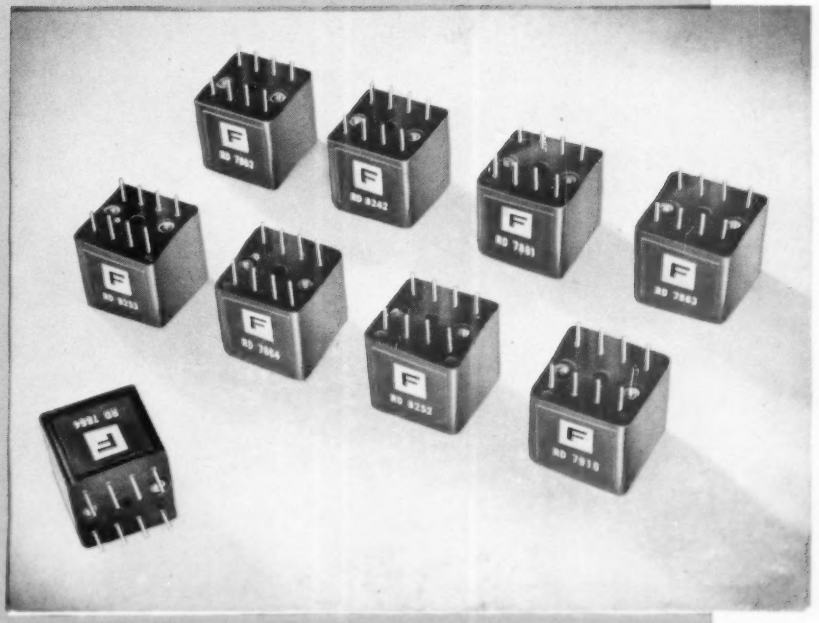

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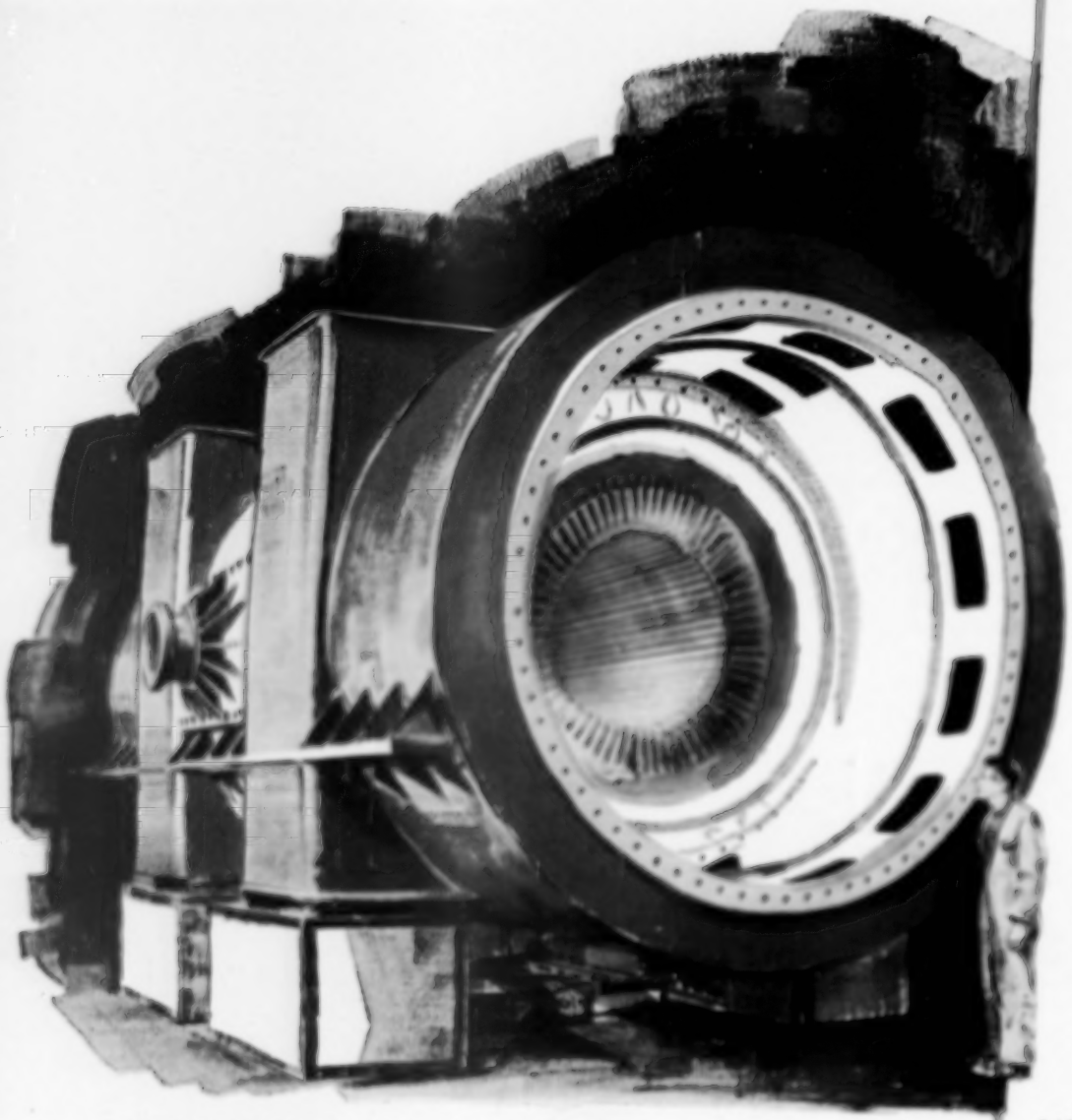

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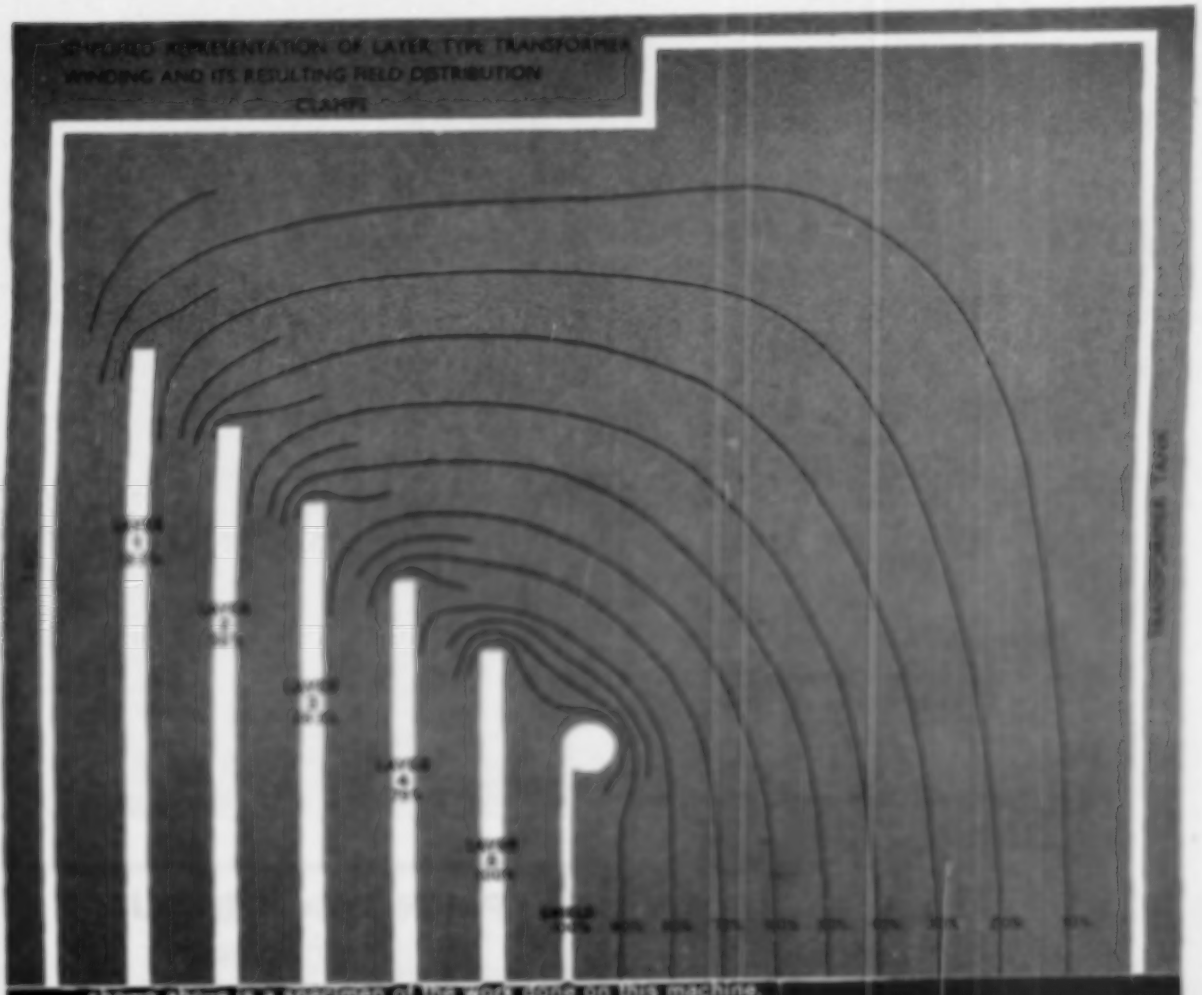

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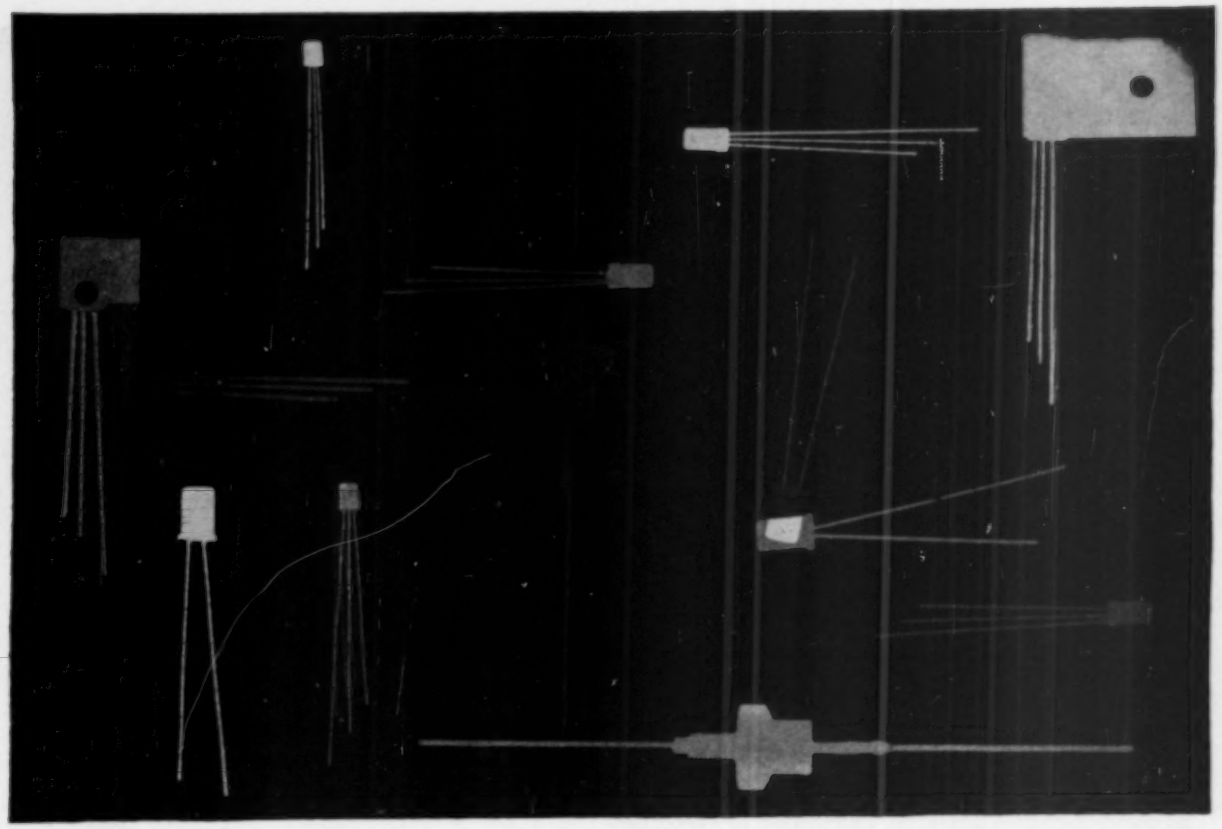

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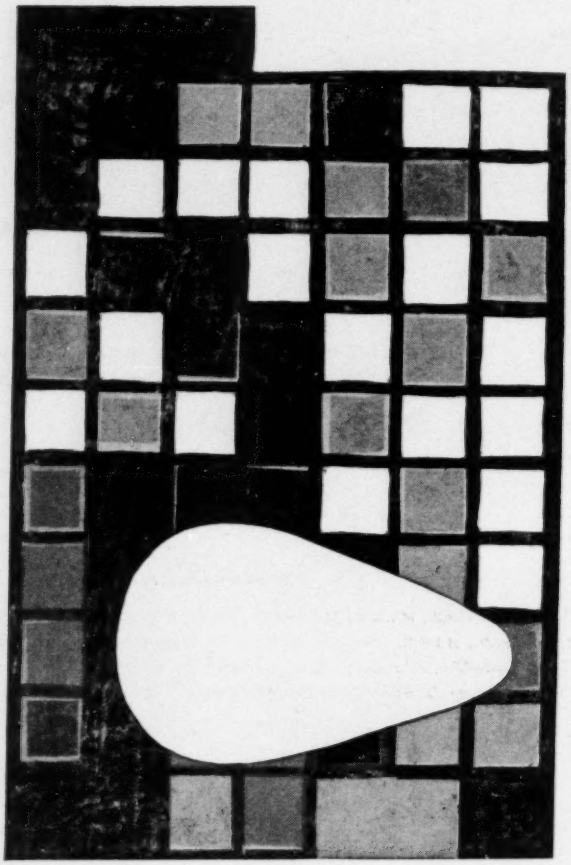

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The Transistor Oscillator

IN recent months our correspondence columns have shown how much interest there is in the question of whether or not the junction transistor has a fundamental mode of operation. Some time ago we said that we doubted if it were possible to answer this question by considering only matters external to the transistor. The correspondence has confirmed our opinion and we regard it as definitely established that it is not possible on circuit grounds to select any one of the three possible modes of operating a junction transistor as being the most fundamental.

As possibly a final demonstration of this, we think that the Hartley form of oscillator is a good example. The junction transistor in the common-emitter connection gives current gain and it has voltage gain when its collector–emitter load equals the base–emitter input resistance. The output and input voltages are in opposite phase. Oscillation will thus occur if a 1 : 1 ratio phase-reversing transformer is connected to feed back the output to the input. In the Hartley oscillator, this transformer is provided by a centre-tapped tuned circuit between base and collector with the tapping joined to the emitter.

In the common-base connection, the current gain is rather less than unity and the voltage gain is under unity also when the collector–base load equals the emitter–base input resistance. The voltages are in the same phase, however. Oscillation can thus be obtained if a tuned circuit is connected between collector and base and a centre-tapping on it taken to the emitter. This acts as an autotransformer so that the collector–base load is four times the emitter–base input impedance. The voltage gain between emitter and collector then approaches four times and, taking the voltage step-down in the transformer into account, the loop gain is nearly two.

In the common-collector connection, the emitter–collector output voltage is necessarily less than the base–collector input, but the phases are the same. For oscillation, a step-up autotransformer is needed. This can be provided by tapping the emitter down a tuned circuit connected between the base and collector.

We have above considered very briefly the transistor in each of its three possible modes of operation in turn and, in each case, the language used to describe the operation is very different. The feedback network must be phase-reversing, non-phase-reversing and voltage step-down, non-phase-reversing and voltage step-up in the three cases. In every case, however, the circuits are identical.

There is only one circuit and it works always in the same way. There are merely three different ways of describing it according to which of the three transistor 'electrodes' we arbitrarily choose as a zero of reference.

Reversible Decimal Counters

USING TRANSISTOR-CIRCUIT TECHNIQUE

By J. L. Goldberg*, B.Sc., B.E.

SUMMARY. *Counters which reverse their direction of counting in response to external control signals are described in this article. These devices are required in the application of optical interferometry to the precise measurement of length.*

Two different scales of notation, namely the binary and the bi-quinary, are shown to be of particular interest in the realization of a reversible decimal scale.

In the binary system, the basis of the device is a scale of sixteen counter formed from four cascaded binary units. With interconnection of the units through two sets of transfer gates, the direction of counting may be reversed by enabling one set and disabling the other.

Two particular numbers in coded binary form are converted to single digits and, by a process of resetting, these effect the necessary corrections to produce a decimal scale in both directions of counting. These corrections are intended to preserve the same constant place weight in both cases and thus make a single decoder necessary to interpret the count at any instant.

A reversible bi-quinary, or 2-5, scale may be realized by interposing transfer gates between a binary stage and a five-element ring counter. No scale correction is required in this case.

Both types of counter are constructed from three basic circuit elements using junction transistors and diodes. The design and use of these elements is considered in detail.

The principles of precise length measurement, using methods of optical interferometry, are now well established[1]. However, the amount of work involved in visual counting of interference fringes is considerable and there is great probability of error both from human causes and from mechanical instability in the interferometer itself. These problems are of particular significance in the case where a 1-metre line standard has to be evaluated absolutely in terms of the wavelength of a spectral line.

At present, the wavelength specified by international agreement for standardization is the orange-red line of the Krypton 86 emission having a value of 6056 Å (1 Å = 10^{-8} cm). The line standard to be measured is arranged as part of a Michelson-type interferometer and, as satisfactory interference is obtainable with Kr 86 to at least 50 cm, this length can be measured as the number of interference lines which appear at a suitable observation point in the optical system when the line standard is moved between reference marks.

The accuracy requirements are of the order of 1 part in 10^8, and since one fringe separation in the interference pattern is 3028 Å the required accuracy represents a measurement to within 1/100 of a fringe separation, and a total count for 50 cm of approximately $1{\cdot}6 \times 10^6$ fringes.

In this article, consideration is given to the problem of suitable electronic-counter designs which can be used for the counting of integral-order fringes.

Because of mechanical instability, it is quite possible for the moving interference pattern to reverse direction for a short distance and then again move forward. The counters are thus required to accept pulses from photoelectric detection circuits and to produce a forward or backward count, depending on the control signals coming from a direction-sensing discriminator. Such a counter is generally known as a reversible counter to distinguish it from another type called a difference counter which produces the difference between two sets of input pulses.

The counting of integral-order fringes requires seven

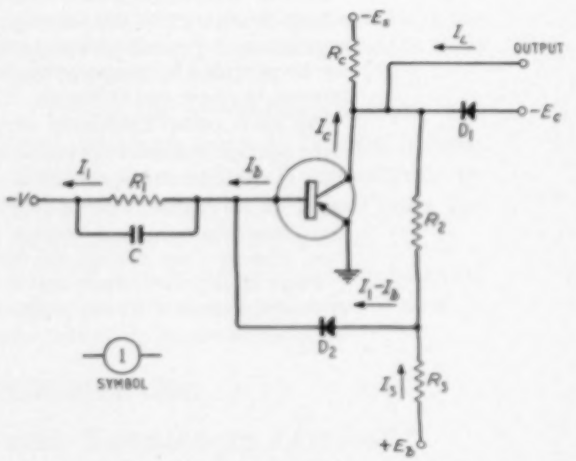


Fig. 1. *Basic circuit element*

* *Division of Metrology, National Standards Laboratory, C.S.I.R.O., Chippendale, N.S.W., Australia.*

decades and it is considered that a resolution of 10 μsec is adequate in this application.

In this article, the design and construction of a single reversible decade is described, since a count of any order of magnitude may be obtained by a serial connection of an appropriate number of decades. The author has constructed a reversible decimal counter by two separate methods, both of which have proved satisfactory. These methods are: (a) A coded binary counter with scale correction for forward and reverse counting; (b) A bi-quinary counter.

Decades by both methods are described below, and in each case they are synthesized from three basic circuit elements:

(i) A voltage-limited transistor amplifier in which the collector-voltage level lies between approximately −0·3 V and −4·5 V with respect to earth and is at one voltage or the other in the stable state (Fig. 1).

(ii) A coincidence gate which has a plurality of inputs and a single output. The input voltage consists of potentials of −0·3 V and/or −4·5 V, corresponding to the collector voltages of a number of amplifiers. If any input voltage is −4·5 V the output is −4·5 V. When, and only when, all the input voltages become −0·3 V, the output voltage rises to −0·3 V (Fig. 9).

(iii) A triggering gate (Fig. 6) which is used for controlling the direction of flow of charge. This is used as a forward or reverse count gate and also, by addition to a simple bistable circuit, it forms a binary counter.

Transistors and diodes are used in the construction except in the display unit where thermionic valves are used to advantage in obtaining a prominent display which may be easily read.

Circuit Design

In the design of transistor trigger circuits, there are certain advantages to be gained by avoiding the use of the saturation region of the characteristic, provided the 'on' or high-conducting condition of the transistor can be otherwise properly defined.

In the saturated or bottomed condition, the current flowing into the base electrode exceeds that which is necessary to maintain the required collector current. An excess of charge is thus needed in the base region to preserve charge neutrality and, with the collector junction forward biased, the average time required for the charge carriers in the base region to diffuse to the collector is many times the average transit time prevailing when the collector is negatively biased with respect to the base (for a p-n-p transistor).

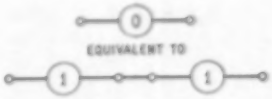
It may be inferred that, for a bottomed transistor, a larger triggering charge than normal must be applied to the base in order to remove the condition of saturation. The longer diffusion time of the charge carriers leads to a delay in the establishment of the opposite, or low-conducting, condition. A satisfactory method of controlling saturation and, at the same time, providing a properly defined 'on' state is by the use of current feedback.

Consider the basic circuit element of Fig. 1, which shows the voltage-limited amplifier previously mentioned. It consists of a common-emitter amplifier with

List of Symbols

α'	= transistor current gain in earthed-emitter condition. (Subscripts max, min refer to upper and lower limits of tolerance for type of transistor used.)
ω_α	= radian cut-off frequency of transistor current gain α
I_{c0}'	= collector current for zero base current in the above condition
R_c	= collector load resistance
I_c	= collector current in the voltage limited regions
I_L	= load current
I_b	= base current
I_i	= input current
e_c	= collector voltage
e_b	= base voltage
E_c	= negative clamp voltage w.r.t. earth
E_s	= negative supply voltage w.r.t. earth
E_b	= positive supply voltage w.r.t. earth
R_d	= forward resistance of diode

voltage limited inverting amplifier



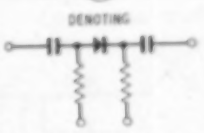
voltage limited non-inverting amplifier

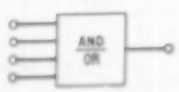
bistable circuit, c, b denote collector and base electrodes

addition gate

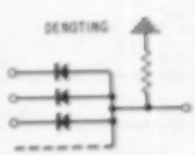
subtraction gate

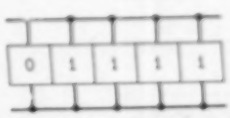
triggering gate





coincidence gate





quinary ring circuit

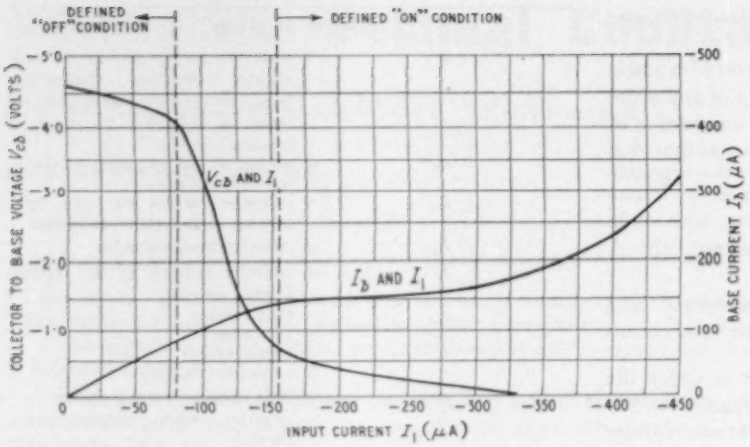


Fig. 2. *Typical static characteristic for a basic circuit element; transistor Type OC71*

two clamping diodes D_1, D_2. With a certain range of input currents, the collector will have two defined levels of voltage.

The more positive level will be attained when the input current defined by $-V/R_1$ is sufficient to bring the collector potential to the level where D_2 conducts, D_1 then being cut off. The less positive level will be reached when the input current is sufficient to cause clamping of the collector voltage at $-E_c$ volts by conduction in D_1, D_2 then being cut off. The current required for the latter condition is clearly less than that required for the establishment of the former condition.

In the first of these states the conduction of D_2, which is made to occur when the collector potential is within a few tenths of a volt negative with respect to the base, provides a feedback current to the base. As the input current is increased in excess of that required to establish the feedback path, this excess current will be provided through D_2 rather than through the transistor-base electrode. Another desirable characteristic is that the conduction of D_2 reduces the current gain (input to output) of the circuit to a fraction of normal. These static circuit characteristics are illustrated for a junction transistor type OC71 in Fig. 2.

The property of an abrupt change in current gain can be used to terminate a transient switching process when the circuit element is used as part of a trigger circuit. Linvill[2] has described circuits in which a Zener diode is used in a similar manner.

In the second state, the input current is sufficient only to cause clamping of the collector at $-E_c$ volts. Conditions are chosen to retain this collector clamp irrespective of external d.c. loading of the output, reasonable variations in temperature and spread of current gain of the transistors. This will ensure adequate definition of the output voltage in the low conduction condition.

Owing to the internal charge always present at a given instant in the base region, the input impedance of the transistor will have a capacitive component. Therefore, a capacitor C is connected in parallel with resistor R_1, providing a compensating step of current to ensure optimum turn-on time.

The basic circuit element of Fig. 1 may be used in

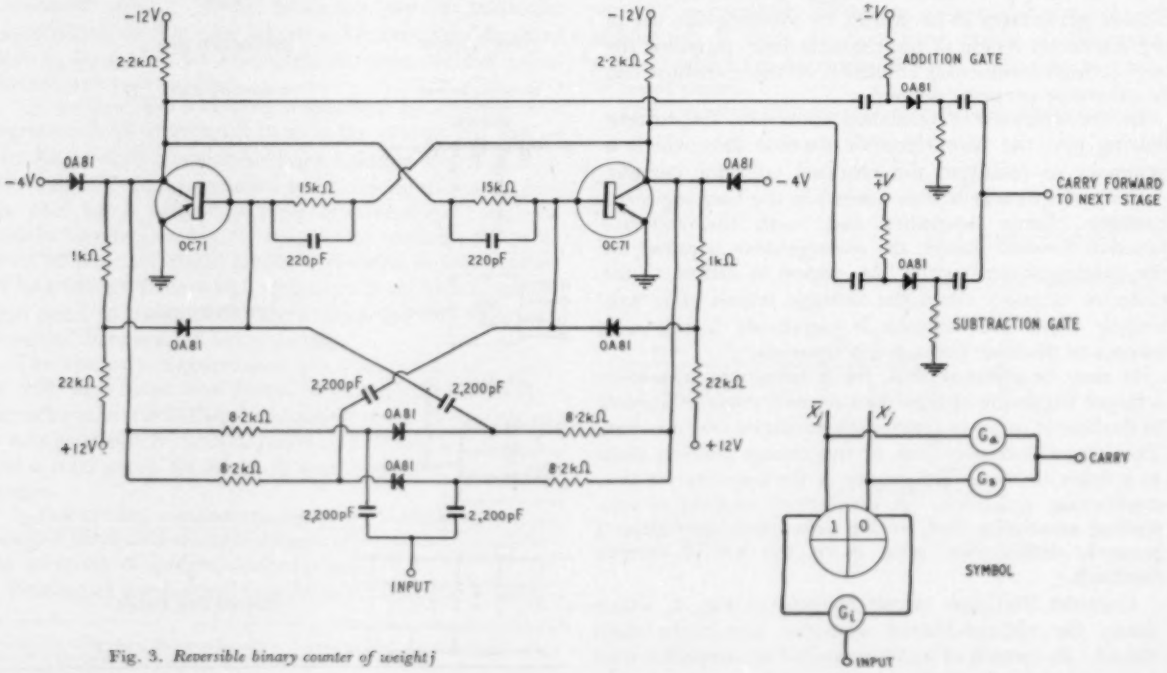


Fig. 3. *Reversible binary counter of weight j*

a number of ways, for example:

(i) Two inverting elements are cross-connected to form a bistable circuit (Fig. 3).

(ii) In binary decades, scale correction is required and two inverting elements are connected in cascade (Fig. 4). This then becomes a non-inverting element and is used to regenerate a waveform with a delay. The delay imposed on the leading edge of the waveform is then approximately $\sqrt{2}\,\tau$ where τ is the rise time of the single inverting element.

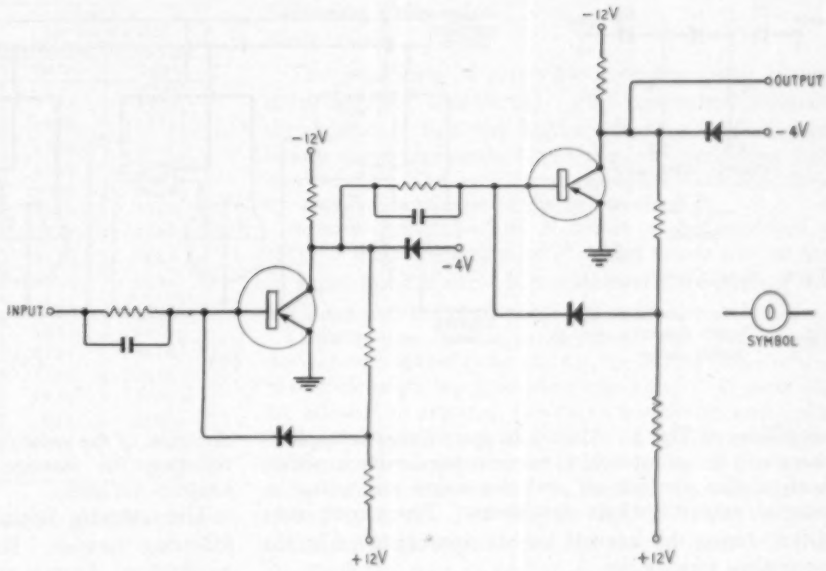


Fig. 4. *Non-inverting delay element*

Derivation of Circuit Operating Conditions (Fig. 1)

High-Conduction Condition

An input current $I_1 \triangleq -V/R_1$ is required to turn on sufficient collector current in order to remove the collector clamp and establish a suitable operating condition with the collector a few tenths of a volt negative with respect to the base.

If $I_{c\,max}$ is the collector current corresponding to this condition (i.e., high conduction) then, in order to turn on this current, a base current is required of at least $I_{c\,max}/\alpha'_{min}$.

If a collector load R_c is chosen so that $I_{c\,max} = E_s/R_c$ then

$$I_1 > E_s/R_c\,\alpha'_{min} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (1)$$

A transistor having $\alpha' = \alpha'_{max}$ may saturate under conditions of inequality [Equ. (1)], but to ensure that the collector-to-base potential is always negative when D_2 is conducting, R_2 and I_3 $(=E_b/R_3)$ are chosen so that $|e_c| - |e_b| < 0$

$$\text{or } (I_1 - I_b)R_d - [I_3 - (I_1 - I_b)]R_2 < 0 \quad .. \quad (2)$$

by applying Kirchhoff's Law, R_d being the forward resistance of the diode. Reasonable variations in I'_{c0} are accounted for by ensuring that $I_{c\,max} \gg I'_{c0}$. When the feedback path is established through the diode, the overall current gain is reduced from α' to a value $(R_d + R_2)/R_c$ approximately (see Appendix 1).

Low-Conduction Condition

For the collector to remain clamped at $-E_c$ volts for the extreme case of $\alpha' = \alpha'_{max}$ requires

$$(I_3 + I_L + I_c)R_c \leqslant |E_s| - |E_c|$$

$$\text{that is} \quad I_b\alpha'_{max} \leqslant \frac{|E_s| - |E_c|}{R_c} - (I_3 + I_L)$$

$$\text{or } I_b = I_1 \leqslant \frac{1}{\alpha'_{max}}\left[\frac{|E_s| - |E_c|}{R_c} - \frac{E_b}{R_2 + R_3} - I_L\right] \quad (3)$$

Here, the effect of I'_{c0} may be included in the value of I_L.

Bistable Circuit

As previously mentioned, two basic circuit elements may be cross-connected to form a bistable circuit. Resistors R_1 are chosen so that with the two defined levels of collector voltage ($-E_c$ volts and 0 volts approximately) the currents I_1 as required by conditions [Equs (1), (2) and (3)] of the previous section are set up in the respective base input circuits. The high- and low-conduction conditions are now mutually maintained. The action of a trigger charge of the required magnitude is to initiate a transient increase of current in one stage which is abruptly terminated by the reduction in current gain of this stage, and is accompanied by a decrease in current of the complementary stage. The transient will grow after triggering provided the circuit is regenerative and this implies a certain minimum for the current gain α'.

Consider a tandem connection of two voltage-limited

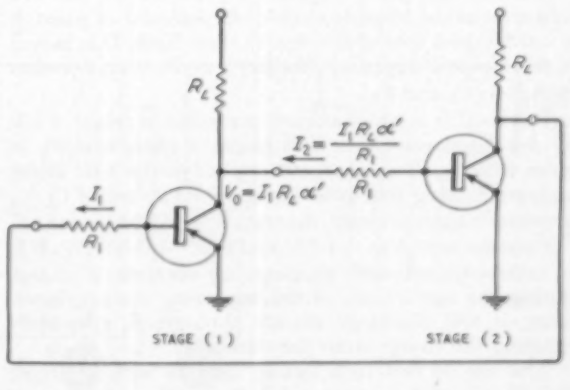


Fig. 5. *Approximate equivalent circuit for regenerative amplifier (non-limited region)*

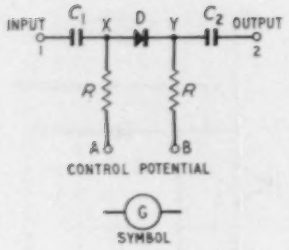


Fig. 6. (*Above*) *Gating circuit* (1)

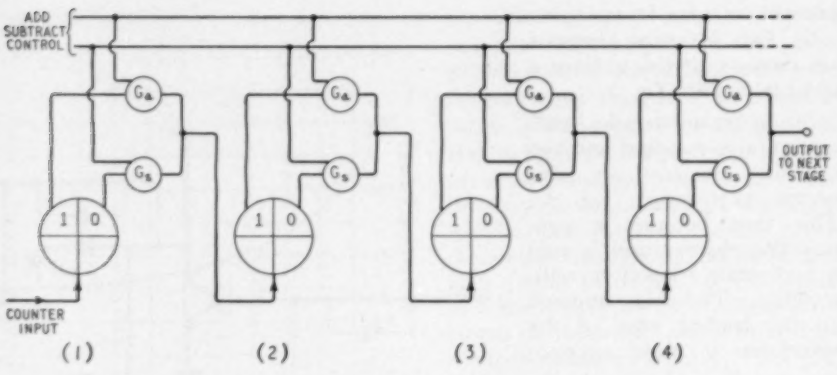


Fig. 7. (*Right*) *Reversible scale of sixteen counter*

amplifiers of Fig. 1. After a trigger charge is applied, there will be an interval of the switching cycle in which both diodes are cut off and the stages are acting as normal earthed-emitter amplifiers. The circuit condition during this interval may be approximated by the equivalent form of Fig. 5.

If I_1 is the input current of the first stage and the voltage output is $I_1\alpha' R_L$ volts, the current supplied to the second stage is approximately $I_1\alpha' R_L/R_1$. If the feedback loop is closed as shown, the output of the second stage supplies the current I_1 to the first.

Thus, the loop current gain is $\dfrac{I_1\alpha' R_L}{R_1}\Big/ I_1$ and the condition for regeneration is that $\alpha' > R_1/R_L$. This condition is easily attained in the designs described.

Gating Circuit Element (1)

The simple bistable element may be triggered, for example, by turning both halves off together by an external stimulus, and the memory of the circuit state is provided by the cross-coupling capacitors which tend to ensure, under certain conditions, that the transient switching process continues in the right direction. However, a gating system is required with a simple bistable circuit in order to ensure reliable operation with inputs which may change their rate from time to time.

The gate used here (Fig. 6) directs a turn-off charge sequentially to either half of the circuit, and provides a memory of its previous condition.

Consider the control potential provided by the two collectors of the bistable circuit. If potential of point A is −0·3 V and that of B −4·5 V, then diode D is biased in the forward direction effecting a connection between capacitors C_1 and C_2.

Terminal 1, initially at earth potential, is raised +4 V by the input waveform and places a charge on C_1 in series with C_2. This charge is applied to the base of the high-conducting transistor via terminal 2, and if C_1, C_2 are chosen appropriately, the transistor will be turned off.

Consider now A at −4·5 V and B at −0·3 V. Diode D is reverse-biased and although C_1 receives a charge during the application of the triggering voltage waveform, it will discharge mainly through R, effectively isolating the charge from the transistors.

The use of two such gating circuits with a simple bistable circuit ensures that the trigger charge is directed to only one transistor at a time, effectively fixing the direction of the switching process and, at the same time, reducing the average loading on the circuit which supplies the gate.

The memory function of the gate occurs in the following manner. The gate is set so that diode D is conducting. As soon as the switching action starts in the bistable circuit due to a trigger charge, A and B rise and fall virtually instantaneously with respect to potentials at X and Y. The gate is thus enabled during the switching process and does not close until after this process has ceased. Thus, capacitors C retain memory of the state of the bistable circuit and prevent triggering via the other gate.

A gate of the same type with add-subtract control potentials is used in order to alter the transfer paths of the carry waveforms.

A complete circuit of a reversible binary counter stage is shown in Fig. 3.

Triggering Requirements for the Bistable Circuit

The circuits are designed to trigger with a 4-V waveform, this being the amplitude available to trigger a subsequent binary stage from a previous stage, when the stages are serially connected.

Linvill[2] gives an approximate relationship for the minimum charge required for switching. A unit charge (1 coulomb) applied to the transistor base will be swept to the collector in approximately one radian period ($1/\omega_0$ sec) and will constitute a collector-current increase of ω_0 amperes, ω_0 being the radian cut-off frequency of the transistor.

The minimum charge required to change the state of the counter is that amount which, if passed through the base of the high-conducting transistor, will bring its collector current to the value possessed by the low-conducting transistor.

Thus, if the difference in collector currents in the two states is I_0, then the charge required is approximately $q = I_0/\omega_0$.

The charge is defined by the capacitors C_1 and C_2 in the gate circuit and the input-voltage step V, thus

$$q = C\,V = \frac{C_1 C_2}{C_1 + C_2}\,V$$

For example, for the circuit conditions of Fig. 3,

$$I_0 = I_{c\ max} - I_{c\ min}$$
$$= 4{\cdot}5\text{mA}$$

	TABLE 1 ADDITION CODE		TABLE 2 SUBTRACTION CODE	
INPUT COUNT	LEFT-HAND STATES	RIGHT-HAND STATES	LEFT-HAND STATES	RIGHT-HAND STATES
0	1111	0000	1111	0000
1	0111	1000	0000	1111
2	1011	0100	1000	0111
3	0011	1100	0100	1011
4	1101	0010	1100	0011
5	0101	1010	0010	1101
6	1001	0110	1010	0101
7	0001	1110	0110	1001
8	1110	0001	1110	0001
9	0110	1001	0001	1110
10	1010	0101	1001	0110
11	0010	1101	0101	1010
12	1100	0011	1101	0010
13	0100	1011	0011	1100
14	1000	0111	1011	0100
15	0000	1111	0111	1000
16	1111	0000	1111	0000

Fig. 8. *Tables 1 and 2 show the sequence of states of the binary elements of Fig. 7 for 16 counts*

$$\omega_{0\ min} = 350{,}000 \times 2\pi \text{ radians/sec}$$

$$q_{min} = \frac{4{\cdot}5 \times 10^{-3}}{0{\cdot}35 \times 2\pi \times 10^{6}}$$

$$= 2 \times 10^{-9} \text{ coulomb.}$$

A safe margin is provided by choosing the capacitors to ensure a charge of twice this magnitude by means of a capacitance of say about 1000 pF (two 2200 pF in series).

With this input charge, average rise-and-fall times of approximately 2 μsec were observed with the bistable circuit of Fig. 3, with cross-coupling capacitances C_1 (= 220 pF) chosen experimentally to make the times equal. The value of C_1 is about one-fifth of that required for a saturated bistable circuit.

The resolution of the counter is determined mainly by the gating-circuit time constants and is about 8–10 μsec with the values adopted.

It has been found, in practice, that the design procedure as outlined in the preceding paragraphs is satisfactory for the production of large numbers of counters, without the necessity for individual trimming adjustments.

Counting Principles

Binary System

The principles of reversible counting using binary units are first considered. The convention adopted throughout is that the high-conducting half of each binary stage represents 1 and the low-conducting half represents 0. The counters are designed to change state by application of positive-going waveforms.

A serial connection of N binary stages produces a radix or scale of notation of 2^N. This means that on the 2^N input pulse, a carry is produced at the output of the last stage and the counter reverts to its initial state[3].

Consider now, an interconnection of four binary stages through two sets of gates G_a, G_s, the initial states of the binary elements being as shown in Fig. 7. If gates G_a are allowed to transmit the carry waveforms and gates G_s are inhibited, then a sequence of 16 counts will set up the states as indicated in Table 1 (Fig. 8). If now, gates G_a are inhibited and G_s enabled, the sequence of states is reversed. This may be checked by noting that, with the convention adopted, the change from 0 to 1 in the left-hand state of one stage produces a change of state in the following stage. The reverse sequence is as shown in Table 2 (Fig. 8).

It is clear that, if a binary scale of 2^4 is acceptable, then reversible counting is a straightforward procedure and consists merely of changing the paths of transmission of the carry waveforms, and weighting factors of 2^0, 2^1, 2^2, 2^3, can be assigned to counters (1), (2), (3), (4), respectively in both directions of counting.

For example, a count of 5 in the forward direction will set the right-hand states to 1010. With the least significant binary digit on the left, the interpretation is

$$2^0 \times 1 + 2^1 \times 0 + 2^2 \times 1 + 2^3 \times 0 = 5$$

Suppose now, the counter is set to subtract, then the next state will be 0010 which is

$$2^0 \times 0 + 2^1 \times 0 + 2^2 \times 1 + 2^3 \times 0 = 4$$

Correction to the Decimal Scale

Correction of a scale of 16 counter (e.g., a four-stage binary to the scale of 10) is usually done in unidirectional counters by taking advantage of the fact that the transition 1110 → 0001 (binary 7 to binary 8) produces a carry waveform in the fourth place (0 → 1) and the transition occurs only once in a count of 16. This

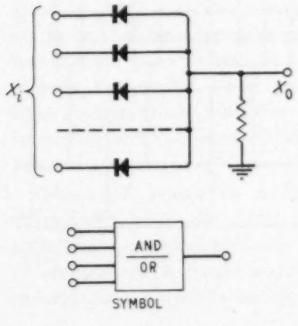


Fig. 9. *AND-OR gate; X_0 is 0 if any or all X_i are 0; X_0 is 1 only if all X_i are 1*

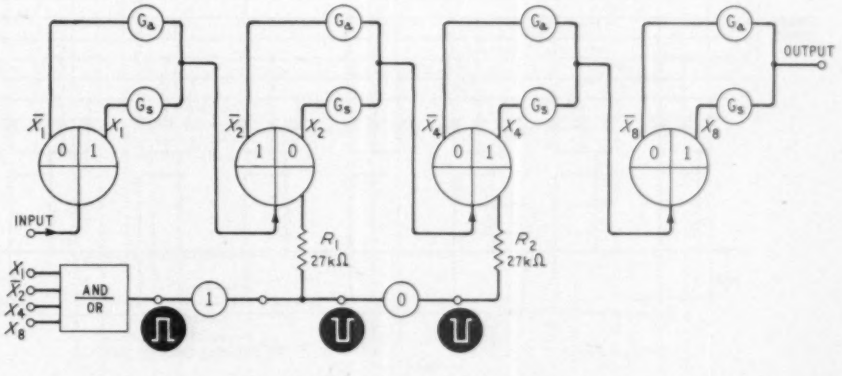


Fig. 10. *The scale-correction subtractive mode*

waveform is used to set the second and third places by feedback and produces the transition 0001 → 0111.

The effect is to produce binary 14 on the count of 8 and binary 15 (1111) on the count of 9. From another point of view, six counts are passed over and effectively subtracted, thus producing a scale of 10.

A further effect is to change the weight or place value of the fourth digit from 2^3 to 2^1. Grossdoff[4] and Slatter[5] describe decimal counters employing this resetting principle.

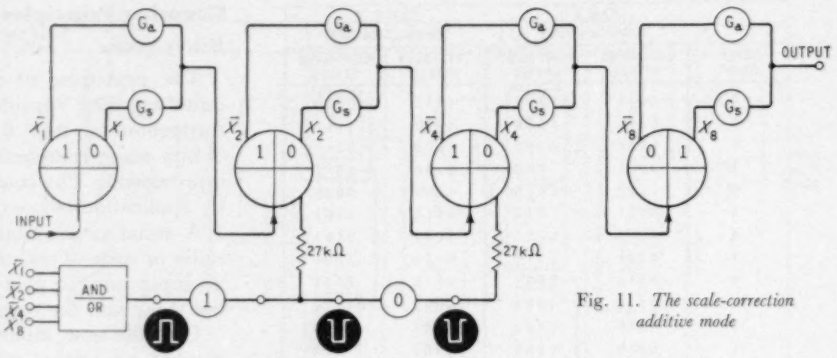


Fig. 11. *The scale-correction additive mode*

An inspection of Tables 1 and 2 will show that this same method, if applied with the subtractive gates in operation, will not produce the required corrections so that the same weighting factors are applicable in both directions.

To overcome this difficulty, the following method of correction, which is more general, is proposed.

The code required in the forward direction is shown marked in the rectangles in Table 1. It is required to obtain the same coded sequence in reverse when subtraction is being carried out. The sequence of states is, therefore, as shown in the rectangles in Table 2, and the weighting factors 1242 will be preserved with this choice.

The method of obtaining these sequences is as follows. In the forward or additive direction binary 8 (0001) or its binary complement (1110) is detected and a single-digit carry pulse is derived by coincidence gating. This digit is fed back to the counter to produce binary 14 (0111). In the reverse or subtractive direction, binary 13 (1011) is detected and the single digit thus derived is fed back to produce binary 7. Thus, in effect, six counts are subtracted in both directions but in a manner which preserves the weighting factors.

Furthermore, it is to be noted that the corrections are mutually exclusive since, if the counter is working additively, the state which produces correction in the subtractive mode never appears in the additive sequence and vice versa.

Alternative Gating Circuit Element

Before considering the actual means used to construct a reversible decade from binary stages, the properties of a second type of gate must be considered.

This element is a d.c.-coupled device consisting of diodes and a resistor with multiple inputs X_i and output X_0 as shown in Fig. 9.

If all the X_i are 1, then X_0 is 1. If, however, any X_i are 0, the output is pulled more negative in potential and prevents any 1 input from appearing in the output.

Further, if all X_i are 0, the output is 0. The device is thus an AND gate to 1s, also an AND gate to 0s only. It will be called an AND/OR gate in this context.

Reversible Decade Counter Using Binary Stages

Four of the reversible binary stages of Fig. 3 are serially connected and corrections applied as outlined in the earlier section (Counting Principles). These scale corrections are performed in such a way that the place weights 1, 2, 4, 2 are preserved in both directions of counting. The resultant coding required is shown in block form in the tables of Fig. 8.

It remains now to explain the mechanism of the

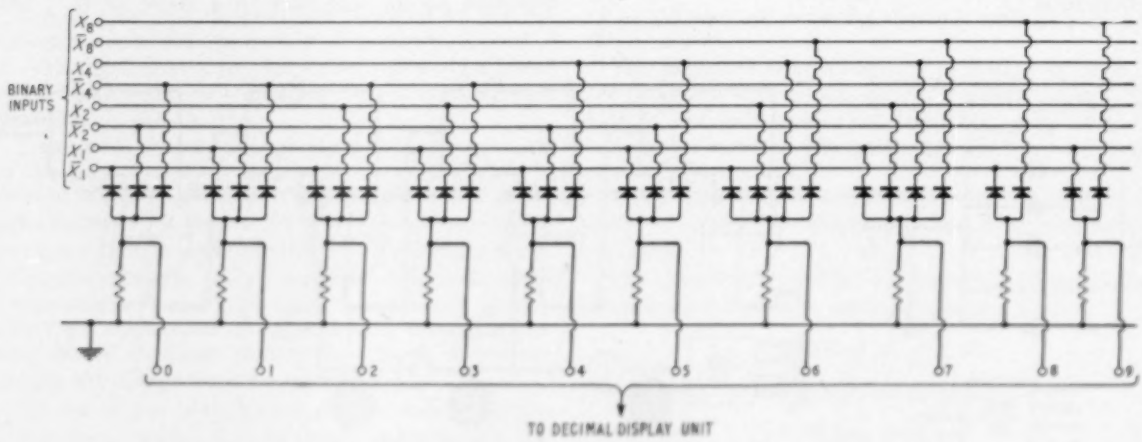


Fig. 12. The binary-to-decimal conversion

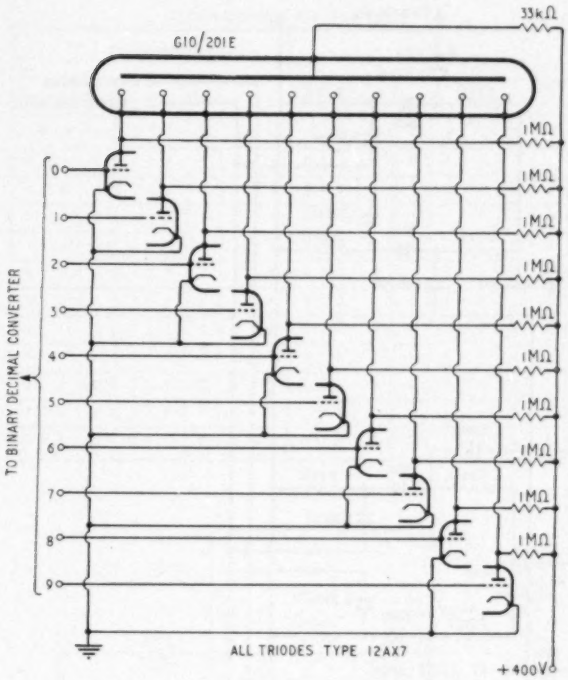


Fig. 13. *A single-digit decimal display unit*

Fig. 14. (*Right*) *Quinary reversible counter*

corrections (Figs. 10, 11). The states of the binary elements as shown are those which occur immediately before correction is applied.

Subtraction Correction (Fig. 10)

The sequence 1011 is detected in an AND/OR gate and produces a positive-going waveform. This is of one digit-period width and is inverted in the standard inverting element and applied to reset X_2. Resistor R, together with the voltage pulse, defines an adequate current which is applied to X_2 base. This causes a transition $1011 \rightarrow 1111 \rightarrow 1101$, since gate G_s is enabled and positive-going waveforms ($0 \rightarrow 1$) cause a transition in the count. The pulse is then regenerated in the non-inverting element and applied to reset X_4.

This will result in a further transition $1101 \rightarrow 1110$. The switching processes by both kinds of element each cause a delay of several μsec. This is necessary to allow the counter to settle down between transitions before further resetting occurs.

Addition Correction (*Fig.* 11)

The sequence 0001 is detected in an AND/OR gate, again producing a positive-going waveform of one digit-period width and the inverted forms reset X_2, X_4 sequentially. The transition $0001 \rightarrow 0111$ is thus effected.

Coded Binary to Decimal Conversion (Fig. 12)

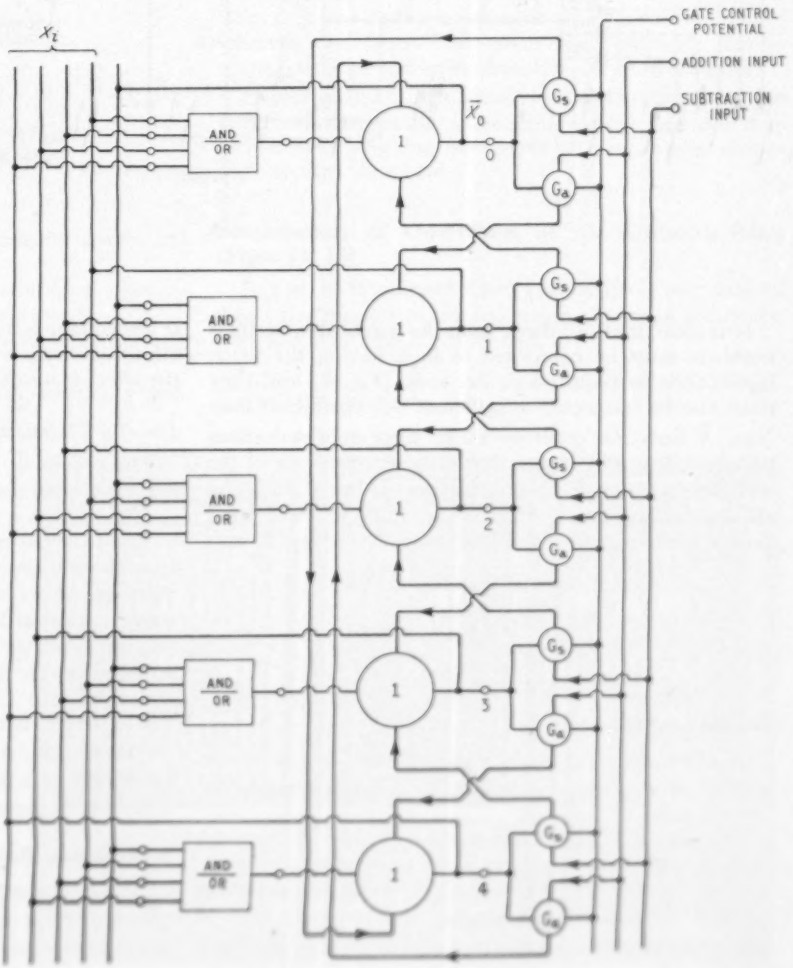
The binary coded form of any decimal number N containing m digits $X_1, X_2 \ldots . X_m$, is

$$N = \sum_{i=1}^{m} X_i 2^{i-1},$$

where for a four-digit code $i = 4$.

A unique pattern of significant digits X_i and their negations $\overline{X_i}$ is generated by the counter for every decimal digit from 0 to 9 inclusive.

Consider a pattern of 1s and 0s applied to an AND/OR gate of the type shown in Fig. 9. If all applied X_i, $\overline{X_i}$ are 1, the gate output will be 1. If any or all X_i, $\overline{X_i}$ are 0, however, the gate output is 0.



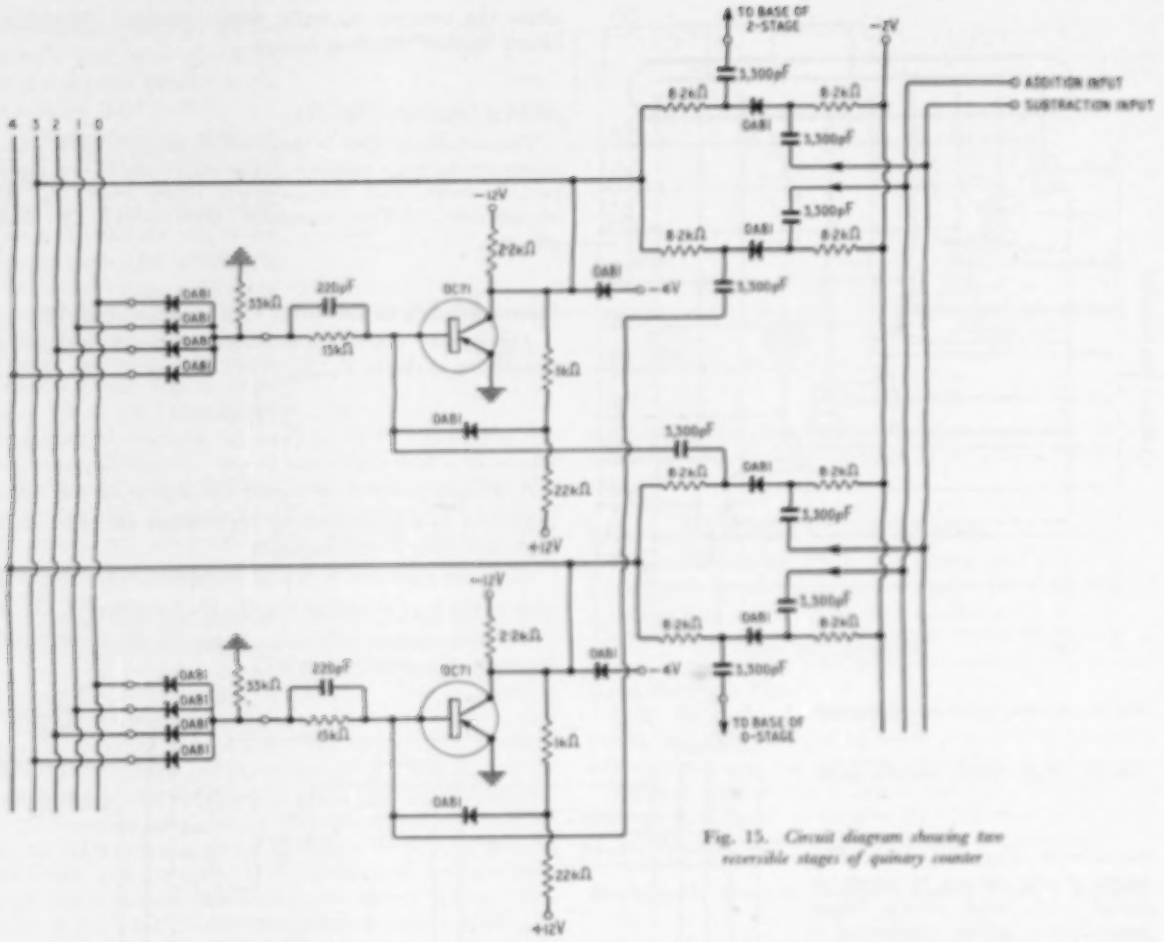


Fig. 15. *Circuit diagram showing two reversible stages of quinary counter*

It is clear that the diode gates for particular decimal numbers must be connected to lines having the place significance as required in the code (Fig. 8) and they must also be connected to sufficient non-significant lines (i.e., $\overline{X}$ lines) in order to make these representations unique. This will ensure that only one gate out of ten will have a 1 output at a given time. After eliminating all redundant inputs, it is found that 30 diodes per decade are required. Examples (taken from Fig. 12) are:

$$2 = (\overline{X}_1, X_2, \overline{X}_4)$$
$$6 = (\overline{X}_1, X_2, X_4, \overline{X}_8)$$
$$8 = (\overline{X}_1, X_8).$$

Numerical Display

A clear in-line display was required and a recent numerical indicator valve G10/201E was available for the purpose.

To use this valve correctly, sufficient differential voltage must be available to cause one numeral to glow and the remainder to be inactive. With transistors at present available in Australia, the required bias of the order of 150 V or more cannot be readily realized. If less voltage is used, an objectionable background from other numerals is superimposed on the wanted one and the effect is unsatisfactory.

Use of a Thermionic Valve as a Switch

The potentials of approximately 0 V and −4·5 V are available from the decoder output and each gate circuit is connected to a separate valve, as shown in Fig. 13.

The valve is used effectively as a switched resistance in each cathode of the neon valve. When grid–cathode volts are about zero, the anode-slope resistance of the valve is of the order of 50 kΩ. In this condition, indicator current is of the order of 3 mA and potential of the cathode of the indicator is about 150 V. The glow stabilizes with 130-V drop, the anode potential being 280 V above earth.

When grid is taken to −4·5 V, valve current is decreased to a small value and anode-slope resistance increases to the extent that the glow extinguishes.

Bi-Quinary Reversible Counter

The bi-quinary 2–5 system, of counting may be used under certain conditions for reversible working. No scale correction is necessary but this is only at the

SEQUENCE OF STATES

FORWARD							REVERSE						
In	Binary		Quinary					In	Binary		Quinary				
	X	X	0	1	2	3	4		X	X	0	1	2	3	4
0	1	0	0	1	1	1	1	0	1	0	0	1	1	1	1
1	0	1	0	1	1	1	1	9	0	1	1	1	1	1	0
2	1	0	1	0	1	1	1	8	1	0	1	1	1	1	0
3	0	1	1	0	1	1	1	7	0	1	1	1	1	0	1
4	1	0	1	1	0	1	1	6	1	0	1	1	1	0	1
5	0	1	1	1	0	1	1	5	0	1	1	1	0	1	1
6	1	0	1	1	1	0	1	4	1	0	1	1	0	1	1
7	0	1	1	1	1	0	1	3	0	1	1	1	0	1	1
8	1	0	1	1	1	1	0	2	1	0	1	1	0	1	1
9	0	1	1	1	1	1	0	1	0	1	0	1	1	1	1
0	1	0	0	1	1	1	1	0	1	0	0	1	1	1	1

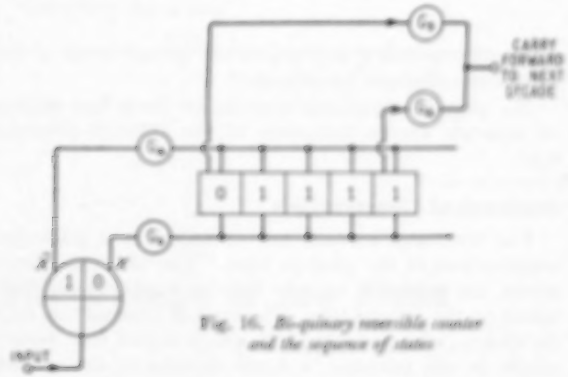


Fig. 16. *Bi-quinary reversible counter and the sequence of states*

expense of considerable internal complexity within the circuits comprising the counter.

A bi-quinary device may be constructed from a binary stage followed by a scale of five-ring counter.

The structure of a quinary ring will now be considered in detail.

An interconnection of five inverting elements and AND/OR gates is made as shown in Fig. 14†. Four of the five elements are interconnected so that each of their respective AND/OR gates receives a 0 input. Under these conditions, the four outputs X_i will all be 1, thus satisfying the condition for the fifth element (element 0) to deliver an output $\overline{X_0} = 0$. There is now one element in a condition of low conduction (output is a 0) and this is, in effect, in a bistable connection with respect to the remaining four which are in a high-conducting condition (outputs are 1). It is clear that the 0 element has control of the 1 elements through one conducting diode for each. Each 1 element is, however, isolated from its neighbour by high reverse-diode resistances.

The d.c. conditions for each 1 element are, therefore, controlled by the 0 element only through the gating system.

A ring configuration of this type will be regenerative provided a certain minimum current gain per stage is available and the switching condition for n similar stages is shown to be $\alpha' > (n-1)R_1/R_e$ by an argument similar to that used in the earlier section for the simple bistable circuit. (See Appendix 2.) This condition in the ring of one 0 and four 1s is preferred to the reverse case of one 1 and four 0s because it is the more statically stable of the two conditions. This may be seen from the fact that the former case will give four elements of low current gain and one element of near normal gain, whereas the latter case gives four elements having appreciable current gain and only one of small gain.

Experimentally, it is found that the condition of the latter arrangement is, in general, astable and this may be checked by setting the system up with inverse circuit logic to that used here.

Mechanism of Operation of the Quinary Ring (Figs. 14, 15)

Just as in the binary case, gates G_a, G_b are used to direct the transfer of the triggering charge in either the subtractive or additive direction.

† Suggested to the author by Dr. M. W. Allen, formerly of C.S.I.R.O. Mathematical Instruments Section.

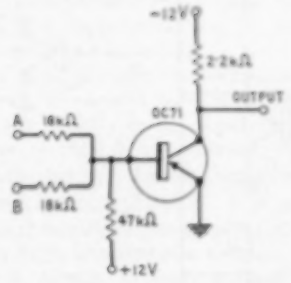


Fig. 17. (*Above*) *Decoder element for bi-quinary counter*

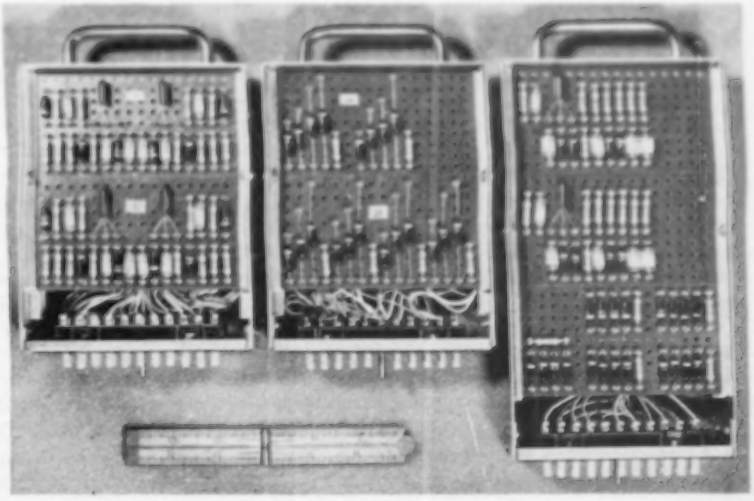


Fig. 18. (*Right*) *Four binary stages in one assembly are shown on the left; a diode decoder in the centre and a quinary ring stage on the right*

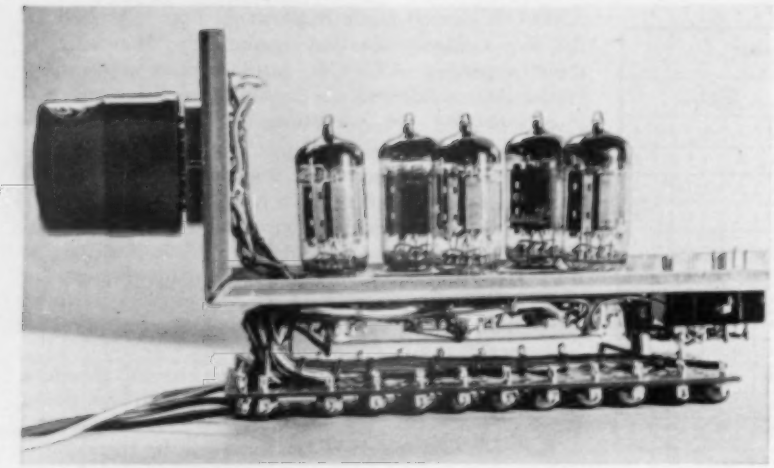
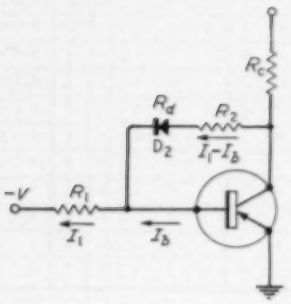


Fig. 19. (*Left*) *Showing the construction of a decade display unit*

Fig. 20. (*Below*) *Equivalent circuit for estimation of current gain in high conducting condition*



Suppose stage 3 is showing a 0; triggering gate G_a will be enabled and a waveform at the addition input will cause a trigger charge to be applied to the base of stage 4; this stage will now show 0 and the circuit logic will ensure that this is the only stage for which this is the case.

It is to be expected from the argument of one of the earlier sections (Counting Principles) that more trigger charge is needed here than for a single binary stage. It is, however, by no means clear what theoretical minimum is required, although it is found experimentally that an increase of about 50% over a simple binary stage is needed.

The subtractive mode of operation follows the same lines as described above and need not be further discussed.

A bi-quinary reversible counter is constructed from the binary counter previously described, together with the quinary ring counter.

By linking together these elements by means of transfer gates G_a, G_s as shown in Fig. 16, a reversible decade is formed, the sequence of states being those described in the table of Fig. 16.

Decoding of Bi-Quinary Count

The 0 state of the binary and the quinary ring counter is detected by means of a circuit which is made to be compatible with the display system already described (Numerical Display). This is shown in Fig. 17, ten such elements being required. In five of them, input A is connected to one side of the binary and each input B to the five individual outputs of the quinary ring. In the remaining five, the input A is connected to the other side of the binary, and B again connected to the quinary ring.

A positive-bias current is fed to the base sufficient to annul the effect of only one 0 input at either A or B. Only when both A and B are 0 will sufficient base current flow to bring the collector potential to near base potential.

Clamping diodes are not necessary in Fig. 17 since the display is only required to capture and hold the final count corresponding to steady-state voltage levels at the transistor collectors concerned.

The stage is compatible with either the in-line display or separate triode indicators of the DM160 (Philips) type.

Mechanical Construction

The transistor circuits are arranged in a packaged construction of the plug-in type. The elements themselves are mounted on one side of matrix board and wired on the reverse side. The type of construction may be clearly seen from Fig. 18 which shows four binary stages in one package, a diode decoder in the second, and the third shows a quinary-ring stage.

Construction of a decade display unit is also shown in Fig. 19.

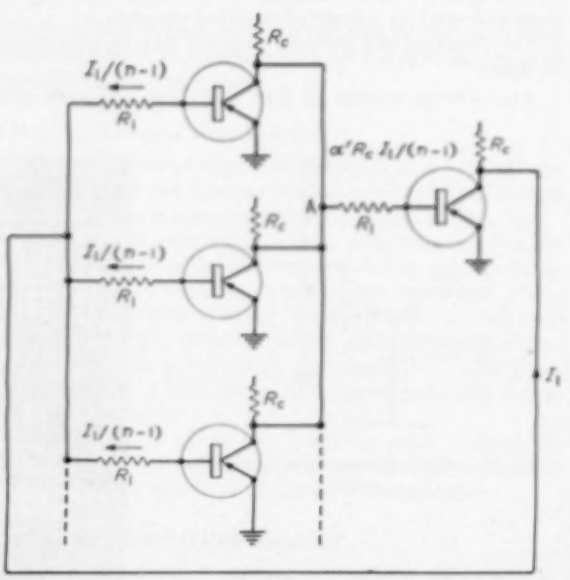


Fig. 21. *Approximate equivalent form for n-stage ring counter as a regenerative amplifier (non-limited condition)*

Acknowledgements

The assistance of Mr. N. J. Marks in the constructional details is gratefully acknowledged.

The work described is part of the research programme of the Division of Metrology, C.S.I.R.O.

APPENDIX 1

In the high-conducting state, diode D_2 conducts and passes a feedback current to the base circuit of amount $I_1 - I_b$ (Fig. 20).

If R_{in} is the input resistance of the transistor

$$I_1 - I_b = \frac{(I_b \alpha' R_c + I_b R_{in})}{R_d + R_2}$$

Thus

$$I_1 = I_b \left[1 + \frac{\alpha' R_c + R_{in}}{R_d + R_2}\right]$$

This equation implies that feedback reduces the current flowing to the base by a factor

$$1 \Big/ \left[1 + \frac{\alpha' R_c + R_{in}}{R_d + R_2}\right]$$

since in the absence of feedback $I_1 = I_b$.

The current gain is thus

$$\frac{\alpha' I_b}{I_1} = \frac{\alpha'}{1 + \dfrac{\alpha' R_c + R_{in}}{R_d + R_2}}$$

With $\alpha' \geqslant 30$, say, $\alpha' R_c \gg R_{in}$ and hence the overall current gain is nearly $\dfrac{R_d + R_2}{R_c}$.

APPENDIX 2

On the assumption that all inverting stages and interconnecting gates are identical, the equivalent circuit of an n-stage ring counter in the process of switching may be simplified to the approximate form shown in Fig. 21.

Suppose the stage to be switched from high conduction to low conduction is on the right of point A; this one stage during transition supplies current to the remaining $(n - 1)$ and if I_1 is the total current, then each stage to the left of A receives $I_1/n - 1$. The voltage available at point A is $\alpha' R_c I_1/n - 1$, and thus the current supplied to the single stage considered is $\alpha' R_c I_1/(n - 1)R_1$, the loop gain is $\alpha' R_c/(n - 1)R_1$ and the circuit is regenerative, provided $\alpha' > (n - 1)\, R_1/R_c$. With $R_1 = 15$ kΩ, $R_c = 2{\cdot}2$ kΩ, $\alpha'_{min} = 30$, a five-element counter is feasible as has been demonstrated.

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The Fringe of the Field ***by Quantum***

THERMIONIC CONVERSION

Simple things that we have always taken for granted often take sudden and unexpected turns for the complex. One very good example is the thermocouple which, now that people have discovered how to tailor thermoelectric materials to need, has blossomed out as the thermoelectric generator; and the theory of its action, too, is probably on its way out of sixth-form syllabuses.

And now there is the thermionic diode, which has suddenly blossomed out—at a time when thermionic tubes are rapidly being displaced by solid-state devices because of their heater consumption—as a heat engine in which this very heater *consumption* is partly converted into useful electrical power. It is not, of course, proposed to heat the cathode electrically; the idea is to make use of very high temperature sources. Nor is it intended to apply a voltage between cathode and anode from an external source; it just has to supply its own. And both electrodes may be plates of considerable area, from square centimetres to square feet. I expect that, during the five or six years over which its possibilities have been investigated, a good deal of technical literature may have come your way. This is merely a short and simple discussion of the principles involved, and an attempt to discover why it works at all.

In the first place, if we have a cathode C at T_1 °K and an anode A at a lower temperature T_2 °K, both maintained at these temperatures by suitably supplying heat to C and causing heat to be rejected from A, we have at least the beginnings of a heat engine with Carnot efficiency $(T_1 - T_2)/T_1$. I suppose it might be possible to develop an analogy to the conventional heat engine with electrons as the working substance, or to treat the diode as a sort of thermocouple with C and A as the junctions, and the electron-stream and the outside circuit as the arms of the couple. But the present approach will be simpler. An electrical generator provides a voltage and circulates a current through the load, in that order. We will take these two partners to the power generated, but in reverse order.

Thermionic Emission: The Current

The important factor determining the emission of electrons from a metal surface is its work-function ϕ, which is the energy that must be supplied to an electron at the top of the Fermi distribution in order to remove it to infinity (Fig. 1). The Richardson equation for the current emitted per unit area by the surface is $I = AT^2 e^{-\phi/kT}$, where ϕ is here expressed in the same units as kT, k is Boltzmann's constant, and A is a factor

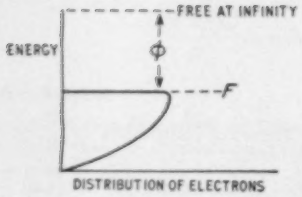


Fig. 1. *Work-function ϕ of a metal represents the energy that must be supplied to an electron at the top of the Fermi distribution to remove it to infinity*

which in theory is a universal constant for all metals, and in practice varies from 162 for caesium to 60 for thoriated tungsten and is as low as 3 for a barium-oxide–strontium-oxide mixture. The value of ϕ in electron volts is 1·8 eV for caesium and 2·6 to 2·8 eV for thoriated tungsten. A low work-function is desirable for the cathode as an emitter; but, as will soon be evident, a high work-function would be useful for the cathode of a generator. The 2·8 eV or so of thoriated tungsten is about the optimum; I use the words "or so" cheerfully, but in fact even the odd 0·1 volt is quite significant. If we take thoriated tungsten at 2000 °K, kT is roughly 0·1725 eV, $e^{-\phi/kT}$ is about 10^{-6}, and the emission works out to about 240 amperes per square centimetre. This temperature is a little above that usually quoted by about 100°; but you can see that with cathode areas of several square centimetres there should be plenty of current available. It now remains to see to the volts.

Work-Function. The Potential Difference

Now suppose that C and A are of different materials, with work-functions ϕ_C and ϕ_A, ϕ_C being greater than ϕ_A. The Fermi levels for these two metals are F_C and F_A respectively. In the first place, forgetting all about current and temperature difference, let us see what will happen if we join C and A in the outside circuit by a conductor which is made up of their two materials, and joins at X. Under these static conditions, the first thing that happens is a transfer of a relatively small number of electrons one way or the other so that the levels F_C and F_A coincide (Fig. 2). This done, as $\phi_C > \phi_A$, remembering that ϕ represents work that has to be *done* on electrons to give them their freedom, the electrons in C at the top of the Fermi distribution have a lower energy than those in A; there is then a transfer of electrons from A to C, so that a potential difference appears at the junction, the contact potential difference $(\phi_C - \phi_A)$ in volts, with A positive to C; this is a p.d. across the diode (Fig. 3). This may be of the order of a volt. So we have the beginnings of a generator which might circulate hundreds of amperes at least, with an e.m.f. of a volt or so. Since, in a working thermionic converter, great emphasis is laid on the need for having ϕ_C in the 2- to 3-volt region and ϕ_A as low as possible, it seems reasonable to accept this over-simplified picture as a first approximation. But it cannot be the whole story when a current is actually circulating. For there is then a continual transference of electrons from A to C, and the initial equating of the Fermi levels is continually being displaced. It is probably nearer the truth to write the e.m.f. as $(F_C + \phi_C) - (F_A + \phi_A)$; but as in practice both electrodes can be made of tungsten (with different work-functions) in many cases it probably doesn't make any difference. What happens in an actual circuit, which isn't made up externally of two kinds of tungsten, will be discussed lower down.

Space-Charge and Other Problems

So far we have been thinking, separately, of currents that are very large, and potential differences that are very small. Usually potential differences of the order of several volts at least are needed to circulate appreciable currents through an ordinary diode, because of the space-charge effect of the cloud of electrons itself, which tends to hang about the cathode repulsively unless it is in a potential gradient sufficiently great to accelerate it quickly out of the way. The well-known Langmuir "three-halves-power law", indeed, starts off with very modest, and modestly-increasing, currents. The remedy, to neutralize the effect of the electron-cloud by a cloud of positive ions, has long been used in gas-filled diodes and mercury-arc rectifiers, devices which must pass large currents if they are to be of any use at all. In thermionic-conversion diodes the gas used is caesium vapour, at a pressure of about 0·2 mm of mercury, and the devices are already known as 'caesium diodes'. Caesium has a very low ionization potential, 3·89 volts; less than half that of mercury vapour and rather more than a quarter of that of argon, and lower than the work-function of pure tungsten, which is about 4·5 eV. I seem, in quoting the tabulated figures, to be mixing up my units; but you will not be confused between an ionization potential in volts and a work-function in electron-volts, while if I don't use the conventionally-expressed data I certainly shall. The point is, that a caesium atom in contact with a pure tungsten surface will be able to spill an electron into a lower-energy sink; and so ionization of the caesium can be done in this way, by making *part* of the cathode of pure tungsten.

Caesium also provides a bonus in the way it acts on the anode material. Positive ions in the neighbourhood of the surface, of oxygenated tungsten, attract electrons by their dipole action, so that the effective work-function of the anode is lowered.

So far then, we have the means of generating a current and maintaining a voltage—much like a simple cell that generates a voltage and maintains a current. In a simple cell, we meet the problems of internal resistance and polarization back e.m.f. In the thermionic cell, they have their counterparts, but *outside* the cell. For (Fig. 4) if the cell passes a current I through an external load, there is an IR drop across the leads; and also, with

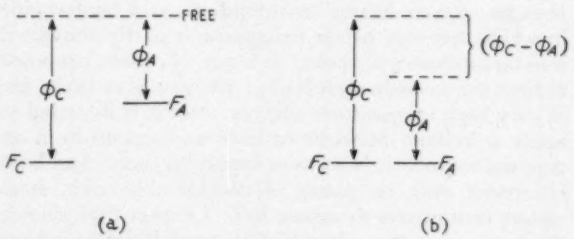


Fig. 2. *(a) Fermi levels of metals A and C, as separate; (b) Fermi levels of metals A and C in contact, adjusted to common level by electron flow; since the energy of an electron in C is lower than in A, electrons flow from A to C through the junction; this causes a p.d. $(\phi_C - \phi_A)$ between the unconnected ends of A and C, with A positive*

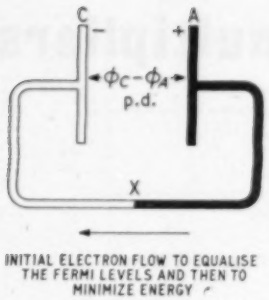


Fig. 3. *Representing what the caption of Fig. 2 stated. This is how the 'voltage' of the thermionic converter arises*

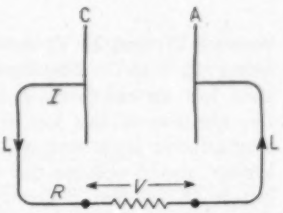


Fig. 4. *On load, the direction of the conventional current is as shown by the arrow. There will be an IR potential drop in the leads, and also a thermoelectric e.m.f., so that (as with cells and other generators) the output voltage is less than the generated e.m.f.*

a 'hot junction' and a 'cold junction' and a temperature gradient round the leads, there is a thermoelectric e.m.f. which could in principle be in either sense, and depends on the thermoelectric power of the leads material and the extreme temperatures. So the *available* voltage will be more like

$$[(F_C + \phi_C) - (F_A + \phi_A) - IR \pm \alpha\,(T_1 - T_2)].$$

Efficiency

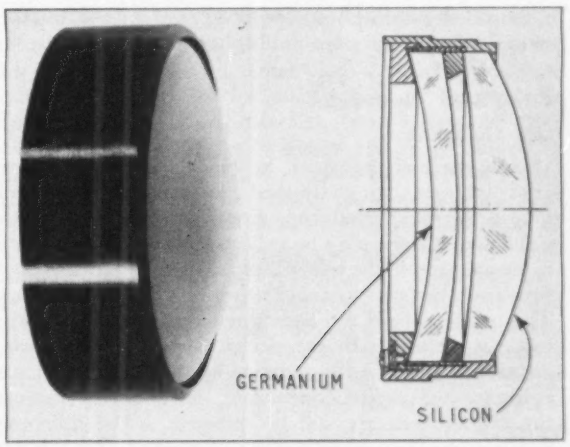
Figures quoted in a recent book, "Thermoelectric Materials and Devices", edited by I. B. Cadoff and E. Miller (Reinhold Publishing Corporation, 1960; in U.K., Chapman & Hall) cover prototypes under development, and calculate possibilities. For the SNAP III-C thermionic generator, which uses a radioisotope source, working between 1400 °K and 800 °K, the efficiency is 5·9%. Of a 60-W heat input from the isotope, 3·54 W is delivered as electrical power (5·42 A at 0·655 V), 32·9 W lost by radiation, 8·36 W lost by thermal conduction, and 15·2 W 'electron cooling loss' taken as a sort of latent heat of vaporization and dissipated. The Carnot efficiency for these two temperatures is 600/1400 = 43%. Design figures for a larger reactor generator, in which the cathode appears to form the outside of a small dustbin-like vessel containing the reactor, speak of an output of 27 kW, temperatures of 1900 °K and 1400 °K, and an efficiency of 9·7% as against a Carnot efficiency of 21%; this, in fact, is the thoriated-tungsten caesium-vapour scheme I have been describing.

It would be an interesting exercise to work out from elementary principles the radiation loss, in terms of Stefan's constant and the fourth-power law, and the electron-cooling loss ($3/2\,kT$ for each electron emitted from the cathode). In fact, only the realization that I have come to the end of my space prevents this. For it seems that both contain a note of warning. The thermionic converter promises to work well at temperatures of the order of 2000 °K. To be sure, we cannot visualize solid surfaces at very considerably higher temperatures than this, but if surfaces above 2000 °K could be used they would have to cope with radiation losses varying as the *fourth power* of the temperature, and electron cooling losses proportional to IT which must increase more steeply than the cube of the absolute temperature. So it may be that the most useful range of temperatures is somewhere near the upper limit at which a reasonable efficiency can be obtained; though I suppose if you can't go working up T_1 you can improve matters by lowering T_2.

In a recent report in *Nature* (6th April 1961), Dr. K. H. Spring of the Central Electricity Board discussed the various means now available for converting heat energy directly into electrical energy, a report based on a series of lectures that some of you may have attended. He mentions there other methods of attacking the space-charge problem, by the use of electric and magnetic fields; but I have come across no details of this so far. A further point that is made in Cadoff and Miller's book is that, with its contents of caesium ions and electrons, the caesium diode is really being operated by a hot plasma, which suggests that the methods now being used in a plasma that is made in two separate halves may perhaps be adaptable to a ready-made hot plasma in some more elaborate machine. In any case, the more conservative among us have one consolation; this is one thermionic application which does not at present seem likely to be superseded by the semiconductors.

EXTENDED-RANGE INFRA-RED LENSES

The production of achromat lenses providing high resolution in the infra-red spectrum has now been extended to include the 8-14



micron range by Servo Corporation of America, 111 New South Road, Hicksville, L.I., N.Y., U.S.A.

DU MONT INTERNATIONAL RENAMED

Fairhill Camera and Instrument Corporation have recently announced the expansion of its overseas marketing organization, with the change of designation of its Du Mont International Division to Fairchild International Division.

Fairchild International will be responsible for the world marketing of the products of other Fairchild divisions and subsidiaries including the products of Allen B Du Mont Laboratories Divisions. Headquarters will continue to be maintained at 515 Madison Avenue, New York, N.Y., U.S.A.

Photoelectric Cells and Photomultipliers

2: MULTIPLIER PHOTOTUBES

By J. Sharpe, B.Sc. A.M.I.E.E.*

The first part of this paper, published in the June issue of *Electronic Technology*, dealt with vacuum photoemissive cells, giving details of the photoemissive surfaces available and considered the problems raised by the necessity of measuring small photocurrents, down to 10^{-16} A. It was pointed out that, although the fundamental limit to accuracy of measurement is set by the statistical fluctuation in the rate of emission of photoelectrons, extraneous sources of noise introduced by measuring apparatus may seriously interfere with the observations.

If some process could be found to multiply each electron by a constant (or nearly constant) factor, the amplitude of the output signal could be raised above the level at which apparatus noise was significant, so that the noise of the device would be reduced to that inherent in the quantized nature of photoemission. Such a process may in fact be realized by making use of the phenomenon of gas multiplication, and also by secondary-emission multiplication. In the former case, a gas-filled photocell is produced and in the latter a secondary-emission photomultiplier tube.

Gas-Filled Photocells

Gas Multiplication

In a gas such as argon, in which free electrons may exist without being trapped to form negative ions, photo-electrons produced from a cathode by the influence of light may be collected by an anode, just as in a vacuum cell, the only difference being that the transit time from cathode to anode is longer with the gas filling. This comes about because the electron is accelerated towards the anode by the electric field, but loses energy on each collision with a gas molecule and since many collisions will occur, depending on the gas pressure, the average velocity will be reduced. (The difference between the two cases may be exemplified by comparing a rocket travelling to the moon with constant acceleration and so with continuously increasing velocity, with a motorist traversing a road with traffic lights every quarter of a mile.)

In a typical gas filling of a few-tenths of a millimetre pressure of argon, the drift velocity of an electron with the values of electric field normally used (between 10 and 100 V/cm) is about 1 cm/μsec. At this gas pressure, an electron emitted from the cathode may achieve an energy sufficient to ionize a gas molecule at one of its collisions when the p.d. between anode and cathode is between 20 and 25 V, the ionization potential of argon being 15·7 eV. The ionized molecule of argon has then lost an electron, so becoming a positive ion and the electron it has lost is free in the gas and indistinguishable from one of the primary electrons. At a higher anode voltage the primary electron may participate in its first ionizing collision far enough away from the anode for it to have a high probability of ionizing another gas molecule in the remainder of its path. In this case, the secondary electron produced from the first interaction also has a good chance of producing ionization before it is collected at the anode, so that now in place of a single photoelectron we have the primary plus three secondary electrons collected at the anode, giving a multiplication factor of 4, while three positive ions are moving towards the cathode. As the anode voltage is increased still further, the number of ionizing collisions made by the primary will increase, owing to the increase in ionizing efficiency with the higher energy gained by the electron between collisions. For n such collisions, the gas multiplication factor M will be 2^n and is found to vary with anode voltage, as shown in Fig. 1.

The slow rise of M above unity is due to the fact that the process described is subject to statistical fluctuation and at values of anode voltage just above the threshold, only a few electrons will make a fruitful collision. For the gas pressures used in photocells, M rises very sharply with anode voltage above a value of 10 to 20, as will be seen.

Movement of Positive Ions

The velocity of a positive ion in a gas under the influence of an electric field is given by $\mu E/p$, where μ is the mobility, which in argon at 1 mm Hg, is about $1{\cdot}3 \times 10^3$ cm^2/volt-sec; E is the electric field, V/cm; and p is the gas pressure in mm Hg. (1 atmosphere = 760 mm Hg.) For a field of 100 V/cm and a gas pressure of $\frac{1}{2}$ mm, the ion velocity is $2{\cdot}6 \times 10^5$ cm/sec,

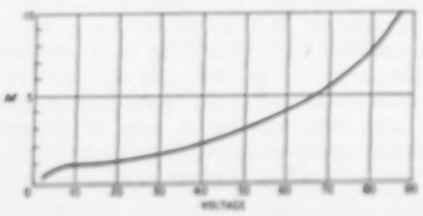


Fig. 1. *Gas multiplication factor M vs voltage for a gas-filled cell*

* E.M.I. Electronics Ltd., Hayes, Middx.

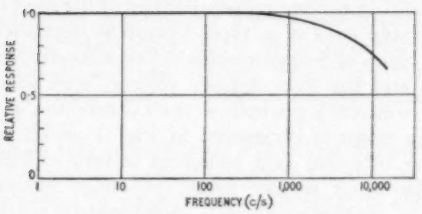


Fig. 2. *Frequency response of gas-filled cell*

so the transit time of an ion over a space of 1 cm is about 4 μsec while the primary electrons cross the same space in the reverse direction in about 1 μsec.

Metastable States and Neutralization of Ions

If the simple picture given above represented the whole story, there would seem no reason for gas-filled cells to respond poorly to light modulated at frequencies well above 100 kc/s. In fact, however, the existence of metastable states in the argon molecule gives rise to complications.

At an electron energy below that capable of causing ionization, the gas molecule may absorb energy from the colliding electron to form an excited state, which has a mean lifetime of about 100 μsec, after which it will emit its energy in the form of a quantum of ultraviolet radiation which may, in turn, release an electron from the tube cathode. Since many of the electron collisions at anode voltages above the threshold for multiplication must result in the production of such excited states, it follows that part of the multiplication is produced by secondary photoelectrons, up to 100 μsec after the initial photoelectron. Many of the metastable molecules, however, will collide with some part of the tube structure before de-excitation by radiation and so transfer their excess energy to the photosurface, with production of an electron. This process may also occur when positive ions are neutralized at the cathode surface.

This course of events causes anode current to flow for a considerable time after cathode illumination is cut off. The current pulse thus formed is terminated by statistical considerations and will be more prolonged the larger the initial electron signal. At sufficiently high excitation the process becomes self maintaining, due to the continuous production of electrons from the cathode by the secondary processes discussed above.

Characteristics of Gas-Filled Photocells

Commercially-available cells are made with AgO.Cs or SbCs surfaces (see Pt. 1, Fig. 4), usually of the opaque type, and are operated up to gas gains of 10 under normal conditions, with values of incident illumination up to about 0·1 lumen. The sensitivity versus voltage varies as in Fig. 1 for the specified incident light levels but, since excessive light on the cathode may cause a self-maintaining discharge which will damage the photosurface, a protective resistance of around 100 kΩ must be included in series with the tube. This introduces nonlinearity due to *IR* drop as the illumination increases, since the gain decreases with the reduction in voltage applied between anode and cathode, but this may be set off by the increase in gas gain with photocurrent. At very low light levels the tube may be used at higher values of gas multiplication, but when dM/dV becomes larger than 0·2 V, the safe limit has been reached.

A typical frequency response for a tube of this type is shown in Fig. 2 and 10,000 c/s is an effective limit. Since, however, the main application for gas-filled cells is in the sound head of film projectors this limitation is acceptable.

The process of gas multiplication fulfils to a first approximation the requirement for 'noiseless' amplification, but the statistical variation in the value of M from electron to electron will increase the noise level above the fundamental limit and although the noise enhancement factor is likely to vary somewhat depending on the tube geometry, the average increase in r.m.s. noise is about 30%.

Where highly-stable characteristics are essential, the gas-filled cell is inferior to the vacuum type, both because of the absence of a plateau in the current–voltage characteristic and because the delicate photoemissive surface is subject to the bombardment of positive ions. Due to the latter effect, gas-filled cells operated at anode currents of several microamperes have a finite life, as compared with the more or less indefinite life to be expected from vacuum cells run under low-current conditions.

Photomultiplier Tubes

Secondary Emission

An electron having an energy of a few hundred electron-volts incident on a surface will communicate this energy to bound electrons close to the solid-vacuum interface and a few of these, having been back-scattered within the material, may approach the interface with sufficient energy to overcome the potential barrier at the surface and emerge as secondary electrons. The number of electrons potentially able to escape from the surface depends on the rate of energy loss of the bombarding particle and the energy communicated to the excited electrons. At low velocity, the incident particle loses energy rapidly but does not give up high energies to the surrounding electrons, while at high incident velocity the communicated energy increases, but the regions of maximum-energy transfer lie deeper into the body of the target material. There should thus be an optimum incident energy at which the maximum value of secondary-emission coefficient δ is produced (δ is the average number of

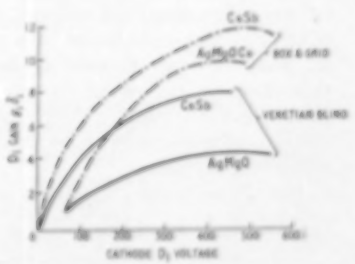


Fig. 3. $g_1\delta_1$ *vs voltage curves for various materials*

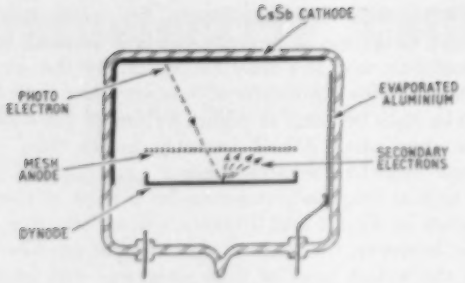


Fig. 4. A single-stage multiplier photocell

secondaries produced per incident electron), as shown in Fig. 3 which gives δ vs voltage curves for materials useful in photomultiplier tubes.

It will be seen that a single stage of secondary emission may be used in a photocell as in Fig. 4 to multiply the photocurrent by a factor of about 5, but the main application is found in multistage tubes, since n stages give a total gain of δ^n.

Secondary-Emission Multipliers

It is convenient to discuss the design of multiplier assemblies separately from the complete cathode-multiplier assembly, since the geometry of the input end of the tube is dependent upon the tube application to a much larger extent than is the multiplying system.

The design problem in a secondary-emission multiplier is to ensure that electrons strike a dynode (secondary-emission elements are called 'dynodes'), at a region where the electric field is directed away from the surface and towards the next dynode and four commonly used geometries are shown in Fig. 5. These are purely electrostatic in operation as contrasted with the early Zworykin multiplier shown in Fig. 6(a) and its derivative in 6 (b), which use crossed magnetic and electric fields, with all the inconvenience in use that this entails.

The electrostatic types of Fig. 5 may be separated into two classes represented by 5 (a) and (b), on the one hand and 5 (c) and (d) on the other, according to the strength of the electric field which accelerates the secondaries from the surface of one dynode. In the venetian-blind [5 (c)], and box and grid [5 (d)] structures, the directing field is comparatively weak and the initial velocity of the secondary electrons plays a predominant part in determining the next landing position, so that there is little correspondence between the point of origin on the parent dynode and the point of arrival on the target dynode and these systems are said to be 'unfocused'*.

The 'focused' structure of 5 (a) and (b) provides a stronger directing field without recourse to grids and so constrains the secondaries into paths with little spread in position from dynode to dynode, with the result that there is less spread in transit time than occurs in the unfocused types. The higher electric fields also allow larger currents to be drawn before space-charge effects supervene.

Not all secondary electrons produced at one dynode are fruitful in producing secondaries at the next dynode, so the stage gain of a given system is dependent both on the value of δ appropriate to the secondary emitting surface and the inter-dynode voltage, and also on the dynode efficiency g which is itself a function of voltage and this point is illustrated in Fig. 7 which indicates that the box and grid structure is very efficient, and with the SbCs surface gives very high values of $g\delta$, while the AgMgO surface is dependent on the presence of caesium to give respectable secondary emission at low incident energy.

The choice of secondary emitting surface is to some extent circumscribed by the type of dynode geometry. Thus, any type of dynode may be made from oxidized silver-magnesium alloy (Ag.MgO), preferential oxidation of the Mg to produce the typical golden colour being carried out before assembly of the system. On the other hand, a precoating of Sb (which would be alloyed with Cs during processing on the pump), tends to suffer from uncontrolled oxidation during handling with consequent variability in secondary-emission properties and better results are obtained by depositing Sb during tube processing from electrically-heated tapes suitably disposed with regard to the dynode surfaces. This technique is more readily applied to unfocused systems and is used in E.M.I. venetian-blind and box and grid tubes.

δ varies approximately as $0{\cdot}2.V^{0{\cdot}7}$ for SbCs where V

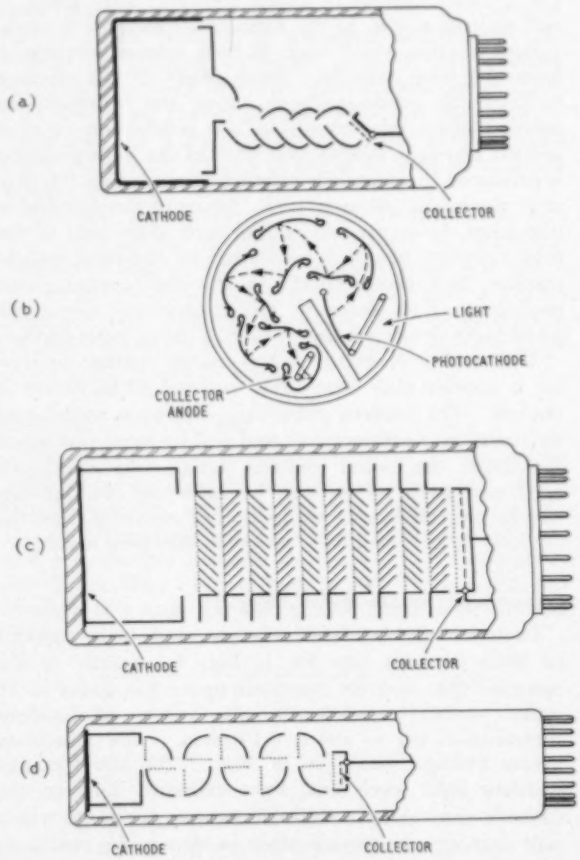


Fig. 5. Electrostatic dynode systems; (a) focused structure, (b) compact focused structure, (c) venetian-blind structure, (d) box-and-grid structure

* For the sake of completeness the Weiss type of multiplier using mesh dynodes, which have low efficiency, and its modern counterpart the transmission-multiplier system used in some image intensifiers, should also be mentioned.

is the interdynode voltage, over the initial part of the curve, while for AgMgO–Cs δ is approximately $V/40$. As a result, tubes made with SbCs material give a higher gain at a lower voltage and a lower rate of change of gain with overall voltage than with AgMgO–Cs.

The usual way of supplying individual dynodes from a divider chain and voltage supply is sketched in Fig. 8 and it is useful to consider the influence on the stage gain of variation in interdynode voltage, which is shown in Fig. 9. It will be seen that each pair of dynodes in the unfocused systems is more or less independent of those on either side, so that quite wide variations from a uniform potential gradient may be applied without unduly interfering with tube performance; the dynode chain thus need not be constructed from high stability resistors. The focused structure is more critical in this respect.

The output section of the multiplier must be designed to cope with relatively high levels of space-charge and

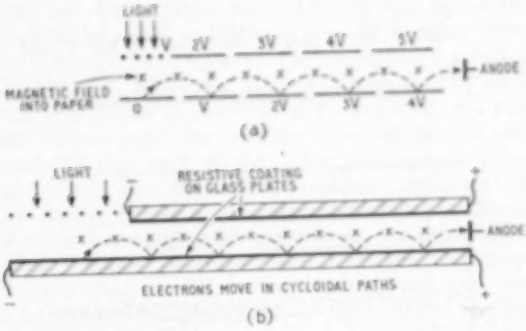


Fig. 6. *Crossed electric and magnetic-field multipliers; (a) Zworykin crossed field multiplier, (b) resistive plate version of (a)*

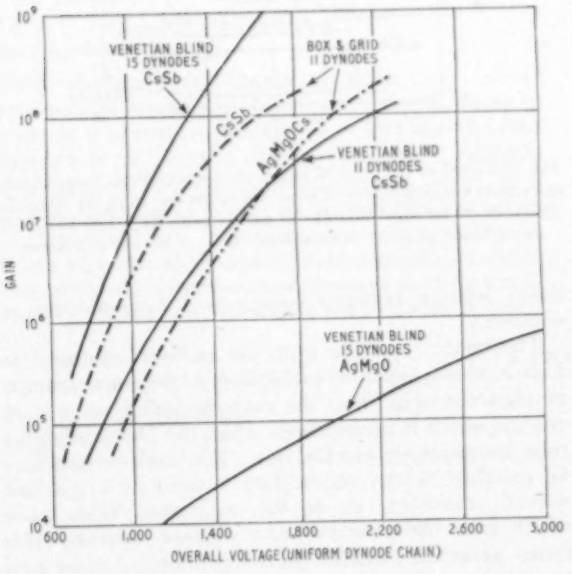


Fig. 7. *Gain vs voltage curves for various dynode systems and secondary emitting surfaces*

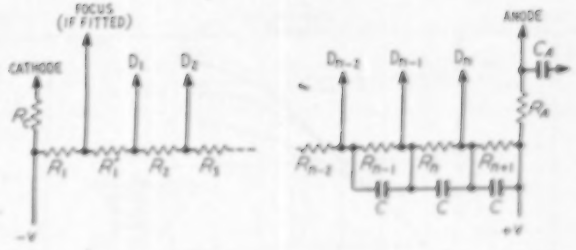


Fig. 8. *Dynode chain parameters. R_c is a protective resistor (100 kΩ); R_1 and R_1' are set to give recommended C–D_1 and C–focus voltage; $R_2 = R_3 = \ldots R_{n-2}$; relative values of R_1, R_2, etc., are set by the overall gain; R_{n-1}, R_n and R_{n+1} may be graded to give interstage voltage appropriate to the required peak current. The dynode resistor current must be at least five times the mean anode current; the value of decoupling capacitor C is appropriate to the a.c. signal; anode load R_A and output capacitance C_A is appropriate to the application*

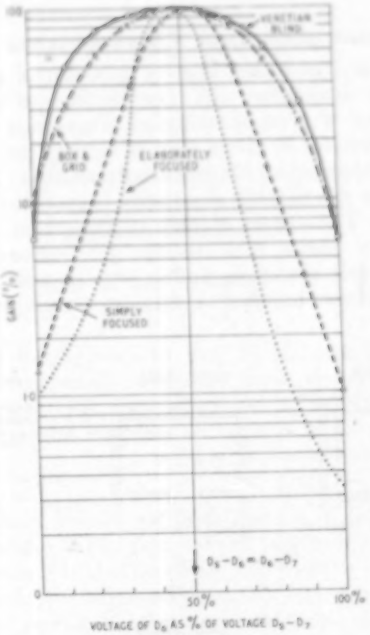


Fig. 9. *Curves showing variation in tube gain as dynode No. 6 is varied in potential between D_5 and D_7*

the anode structure in general consists of a grid through which electrons from the penultimate dynode are directed on to a plane electrode, which forms the last dynode. From Fig. 10 it will be seen that, given high interstage voltage at the output end of the tube, high peak currents may be drawn, although average currents not exceeding 1 mA are normally allowed.

Cathode-First Dynode Systems

The input into the multiplying system must take photoelectrons from the cathode and direct them into the first dynode with the maximum collection efficiency and various structures are shown in Fig. 11 for end-window tubes with semi-transparent cathodes. The compact structure of Fig. 5 (b) is one example of an

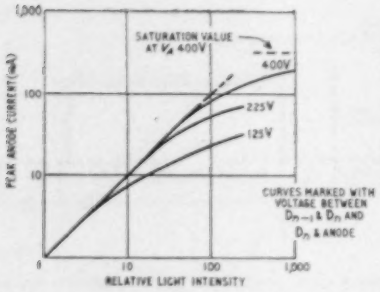


Fig. 10. *Anode saturation characteristics (venetian-blind dynode and anode system)*

opaque cathode design. It will be seen that large area dynodes, such as are offered by the venetian-blind structure are likely to give more efficient and uniform electron collection.

Photomultiplier Tube Characteristics

Just as a photocell may be specified in terms of the cathode sensitivity, μA per lumen, so the overall sensitivity of a photomultiplier is specified in amperes per lumen for a given overall voltage, or better, the voltage required to give a specified sensitivity. In addition, of course the cathode sensitivity is specified in μA per lumen. The curves of Fig. 12 show how the overall sensitivity varies with voltage for various structures and dynode materials and it will be seen that tubes with a larger number of dynodes give a more nearly

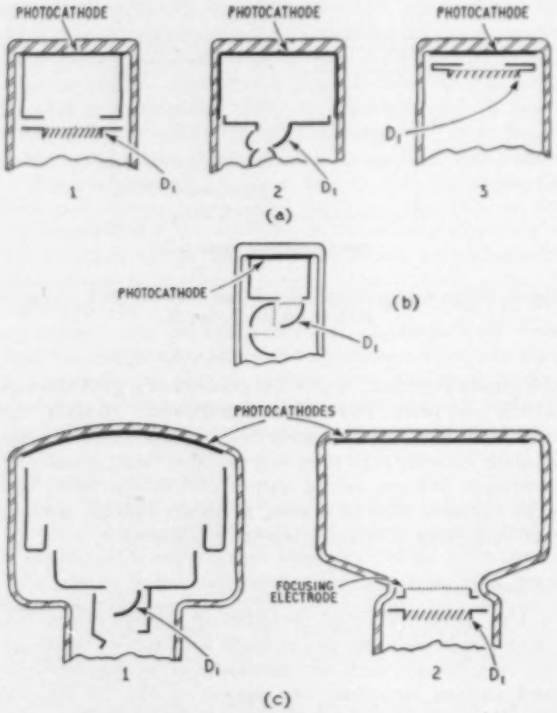


Fig. 11. *Cathode D_1 systems*

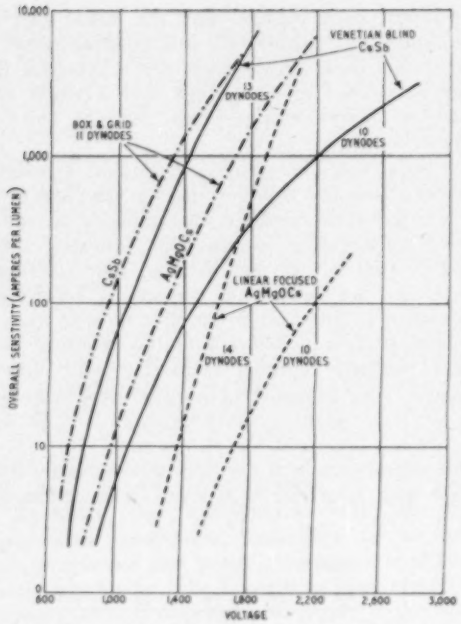


Fig. 12. *Overall sensitivity vs overall voltage*

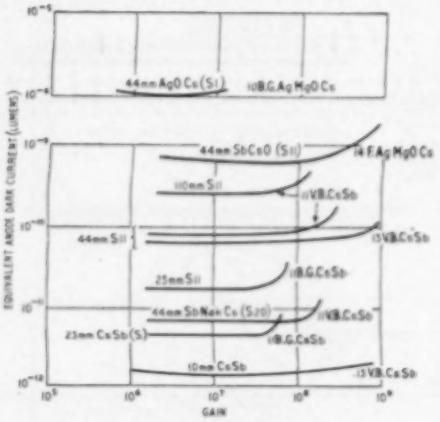


Fig. 13. *Dark current at 23 °C vs gain for various cathode sizes and types and various dynode structures. Typical sensitivities are: S11, 60–70 μA/L.; SbCs, 40–60 μA/L.; S1, 15–30 μA/L, and S20, 120–140 μA/L. F = focused, V.B. = venetian blind, B.G. = box and grid dynodes*

linear relation between voltage and the logarithm of sensitivity.

Thermionic emission from the cathode, discussed in Part 1 of the paper, is multiplied in the same manner as photoelectrons from the cathode and gives a dark current which is present even when the photomultiplier tube is completely blacked out. This dark current may be specified as the anode dark current at a specified overall sensitivity, or as the equivalent light input which gives the observed value of dark current. This latter parameter should be independent of tube gain provided no feedback occurs from anode to cathode, but as shown in Fig. 13, above a certain value of gain,

TABLE 1

C–D_1 System	Cathode Type	Cathode Dia. mm.	Type of Dynode	Dynode Surface	No. of Dynodes	Typical Limiting Gain
11 (a3)	S–11	10	V.B.	SbCs	13	10^8
11 (a3)	S–11	10	V.B.	SbCs	15	2×10^8
11 (a1)	S–11	44	V.B.	SbCs	11	2×10^8
11 (b)	S–11	23	B.G.	SbCs	11	6×10^7
11 (c1)	S–11	44	F.	AgMgOCs	14	5×10^8

depending on the tube type, the equivalent input dark current begins to rise and if the voltage is increased to a sufficiently high value the tube may become unstable. The superproportional rise in anode dark current with gain is due to optical and ionic feedback from the anode to the cathode of the tube and sets in at low gains where the structure provides a more or less open path between input and output. The box and grid structure, with its open axial path, is found to have a value of limiting gain more or less independent of the number of stages, while the more opaque venetian-blind structure has a higher value of limiting gain before feedback sets in and this is increased by a factor of two for each additional pair of dynodes. The dark current is of course dependent on the cathode area and Table 1 shows that the limiting value of gain is also affected by this part of the tube, the highest value being given by the small cathode structure of Fig. 11 (a3). This design also gives the lowest value of dark current per unit area of cathode, since the plane parallel geometry ensures that thermionic electrons from parts of the C–D_1 region which do not play any part in the fruitful absorption of light, will not enter the first dynode. For some reason not yet fully understood, the linear focused structure using AgMgOCs dynodes, gives a dark current about 100 times worse than the venetian-blind system using SbCs dynodes, for S11 cathodes.

Time Characteristics

Each electron leaving the cathode and giving secondary electrons from D_1 gives a pulse of charge at the anode of average amplitude Ge, where G is the tube gain and e is the electron charge, $1{\cdot}6 \times 10^{-19}$ coulomb. For a gain of 10^8, the electron pulse is $1{\cdot}6 \times 10^{-11}$ coulomb, sufficient to charge up a capacitance of 50 pF to a voltage of 1/3 V. The electrons forming this pulse will not all arrive at the anode simultaneously, however, owing to the time dispersion of electrons in their transit through the tube and the anode current will be found

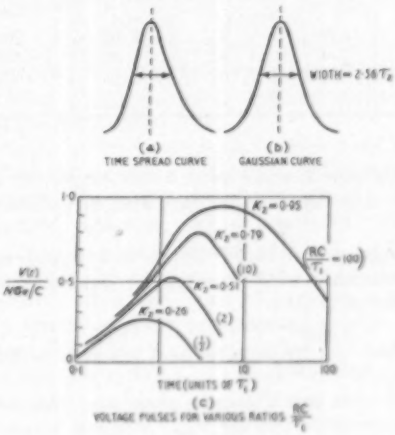


Fig. 14. Time characteristics

to have the form shown in Fig. 14 (a) and although this rises more rapidly and falls more slowly than a Gaussian curve [Fig. 14 (b)], nevertheless it is often convenient for the purposes of calculation to assign a standard deviation $\tau_2 = \left(\frac{\textit{width of pulse at half amplitude}}{2{\cdot}36}\right)$ on the basis of which the peak current is $Ge/\tau_2\sqrt{2\pi}$. For the example already quoted this becomes $6{\cdot}4/\tau_2$ mA if τ_2 is measured in nanoseconds. (10^{-9} sec.)

If, in place of a single electron, the input consists of N electrons produced simultaneously over the whole cathode, the time spread due to the multiplier will be increased by the transit time spread in the C–D_1 space, to give the values of τ_2 given in Table 2.

A total of N photoelectrons produced by an exponentially decaying source of time constant τ_1 (such as a phosphor excited by a nuclear particle) will give a cathode current of $\frac{Ne}{\tau_1}\exp\left(-\frac{t}{\tau_1}\right)$, as a

TABLE 2

C–D_1 System	Cathode Dia. mm.	Type of Dynode	No. of Dynodes	Overall Voltage	Linear Output Current mA	τ_2 nanosec	Gain
11 (a)	44	V.B.	11	2,400	50	6	10^8
11 (b)	23	B.G.	11	1,500	10	8	5×10^7
11	110	V.B.	11	3,000	70	12	2×10^8
11	44	F.	14	2,400	200	1	10^8

V.B. = Venetian blind; B.G. = Box and grid; F. = Linearly focused.

TABLE 3

τ_1/τ_2	Large	4	2	1
K	1	0·65	0·48	0·31

TABLE 4

f	$g_1\delta_1$	$g\delta$	a	C–D_1 Voltage
0·90	6	4	1·16	200 V
0·85	4	4	1·29	100 V
0·80	3	3	1·36	75 V

function of time t, and a peak anode current of $KNGe/\tau_1$ where K is a function of the ratio τ_1/τ_2 tabulated in Table 3.

The voltage produced by such a signal across an anode resistance of R shunted by capacitance C may be calculated to be:

$$V = \frac{NGeR}{\tau_1 - RC}[\exp(-t/\tau_1) - \exp(-t/RC)]$$

provided τ_1 is much greater than τ_2. This expression gives a maximum voltage V_{max} at time

$$t_{max} = \frac{\tau_1}{\tau_1/RC - 1}\log_e(\tau_1/RC).$$ [See Fig. 14 (c)],

where $V_{max} = \frac{NGe}{C}K_2$.

The output voltage across RC due to N electrons produced simultaneously at the cathode is $[NGe/\tau_2\sqrt{2\pi}]\,[AR/(RC/\tau_2)]$ where A is 0·85 for $RC = \tau_2$ and 1·7 for $RC = 5\tau_2$.

The time spread in a photomultiplier tube sets an upper limit to the frequency response, which may be taken to be approximately $1/\pi\tau_2$ c/s., so that a venetian-blind dynode tube with $\tau_2 = 6$ nanoseconds has an upper frequency limit of about 50 Mc/s. At the lower end of the frequency spectrum the photomultiplier tube responds without loss at zero frequency, since the secondary-emission multiplier is, in effect, a linear d.c. amplifier for small input currents.

Noise in Photomultiplier Tubes

It was shown in the last section of the first part of this paper that the relative r.m.s. fluctuations, due to statistics of photoelectron emission, in a charge it made up from a photocurrent i collected over a time t, is equal to $(e/it)^{\frac{1}{2}}$, while the r.m.s. value of noise current measured in a circuit of bandwidth Δf is $(2ei\Delta f)^{\frac{1}{2}}$. The multiplication produced by the secondary-emission process cannot of course reduce this fundamental noise component, so that for a gain G, the minimum output r.m.s. noise current would be $G(2ei\Delta f)^{\frac{1}{2}}$. For an anode current of $I = Gi$ and with an anode resistor R, the total r.m.s. noise voltage is $(eR\Delta f)^{\frac{1}{2}}(2G^2Ri + 0{\cdot}1)^{\frac{1}{2}} = (eR\Delta f)^{\frac{1}{2}}(2GIR + 0{\cdot}1)^{\frac{1}{2}}$, so that RI must exceed $0{\cdot}05/G$ volts if Johnson noise is not to predominate and this is readily achieved.

The statistical variation in the value of δ, the coefficient of secondary emission, at the first dynode of a multiplier increases noise level above the minimum value and it can be shown that the enhancement factor a is equal to

$$f^{-\frac{1}{2}}\left(1 + \frac{1}{g_1\delta_1} + \frac{1}{g_1\delta_1(g\delta - 1)}\right)^{\frac{1}{2}},$$

where f is the proportion of photoelectrons reaching D_1; $g_1\delta_1$ is the D_1 gain and $g\delta$ is the average gain of succeeding stages of the multiplier. Typical values of f, $g_1\delta_1$ and $g\delta$ are given in Table 4, together with the appropriate values of a, and an indication of the cathode–D_1 voltage needed to give the particular values of D_1 gain for a 50-mm diameter venetian-blind SbCs dynode tube. It will be seen that unduly low values of the latter parameter may give rise to serious degradation of the signal to noise properties and it is important to keep the C–D_1 voltage up to the recommended value for the particular tube.

Measurements with P.M. Tubes

From the spectral sensitivity curves given in Part 1 and the values of dark current given appropriate to particular tubes, the data of Table 5 have been calculated to give the input power at a particular wavelength to give a current equal to the dark current, for typical tubes.

These figures give a first-order idea of the sensitivity of a tube for a particular low light-level application when operated at room temperature. When the dark

TABLE 5

Cathode (Type and Size)	Cathode Dark Current amps. (23°)	Wavelength, μ									
		0·2	0·25	0·3	0·35	0·4	0·5	0·6	0·7	0·8	1·0	1·1
		[Power in $\mu\mu$W (10^{-12} Watt)]										
† 10 mm SbCs ..	$2{\cdot}5 \times 10^{-17}$	0·001	0·001	0·0007	0·0006	0·0006	0·001	0·01				
‡ 23 mm BiAgOCs ..	4×10^{-15}	2·5	0·8	0·5	0·35	0·25	0·25	0·63	2·5	2·5		
† 44 mm SbCsO ..	6×10^{-15}	·25	·24	·17	·13	·12	0·2	0·9	1·00			
† 44 mm SbNaKCs ..	$1{\cdot}5 \times 10^{-15}$	·046	·03	·027	·03	·026	·028	·047	0·1	0·75		
		[Power in μW (10^{-6} Watt)]										
§ 44 mm AgO–Cs ..	10^{-10}				0·035	0·05	0·15	0·05	0·03	0·03	0·1	0·4

§ Glass window. † Fused quartz windows. ‡ U.V. transmitting glass window.

current is too high for a critical application, refrigeration may be used to reduce the thermionic emission from the cathode (Fig. 15).

For such applications, when the tube dark current may limit the accuracy of a measurement, the statistical considerations outlined in the first part of a Section (Measurement of Photocurrent) of the previous paper would apply. In that section, example (a) considered the observation of a relatively faint star producing an illumination of 10^{-12} lumen. With a photomultiplier tube having a 10 mm SbCs cathode [type 11(a3)] and an equivalent dark current of $1{\cdot}5 \times 10^{-12}$ lumen and operating at a gain $G = 3 \times 10^7$ (or an overall sensitivity of 2,000 amps per lumen) (e.g. E.M.I. type 9502B), the signal plus dark current, $(I + I_d)$, would be 5×10^{-9} A and I_d would be 3×10^{-9} A. These currents could conveniently be measured by charging up a capacitance of 10,000 pF for a time T sec to give a voltage of the order $\dfrac{5 \times 10^{-9}}{10{,}000 \times 10^{-12}}\ T = 0{\cdot}5\ T$ volts. For linearity, $0{\cdot}5\ T$ volts must be rather less than the anode to last dynode voltage of the photomultiplier tube (say 100 V) and for T of the order of 60 sec, a convenient potential around 30 V would be produced, which is readily measurable with considerable accuracy.

The signal current, deduced by subtraction of the two voltages,

$$C\left[\frac{(I + I_d)}{C}\,t - \frac{I_d t_d}{C}\right],$$

would have a standard error of:

$$\Delta I = \left(\frac{I_d}{Gt_d}\,e + \frac{(I + I_d)}{Gt}\,e\right)^{\frac{1}{2}} G \quad \text{(the cathode current concerned is } i/G\text{)}$$

For $t = t_d = T/2$, $\dfrac{\Delta I}{I} = \left(\dfrac{2Ge}{T}\,\dfrac{(I + 2I_d)}{I^2}\right)^{\frac{1}{2}}$

and for $I = I_d$

$$\frac{\Delta I}{I} = \left(\frac{6Ge}{IT}\right)^{\frac{1}{2}} \simeq \left(\frac{6 \times 3 \times 10^7 \times 1{\cdot}6 \times 10^{-19}}{60 \times 2 \times 10^{-9}}\right)^{\frac{1}{2}} \simeq {\cdot}012$$

so that by use of a photomultiplier tube in place of a photocell we have obtained a precision of about 1% in 60 sec in place of the 10 hr required by a photocell having a dark current of 2×10^{-14} A. This is a reflection of the fact that the latter figure is set more by fluctuating leakage currents in the cell than by true thermionic emission. The comparable figure in the photomultiplier tube is $\dfrac{3 \times 10^{-9}}{3 \times 10^7} = 10^{-16}$ A. (We may quickly check the above calculation as follows: 10^{-16} A = 700 electrons per second. In 30 sec we have 21,000 electrons and the fractional standard error involved in a difference measurement would be: $\dfrac{(42{,}000)^{\frac{1}{2}}}{21{,}000} \simeq \dfrac{205}{21{,}000} \simeq {\cdot}01$)

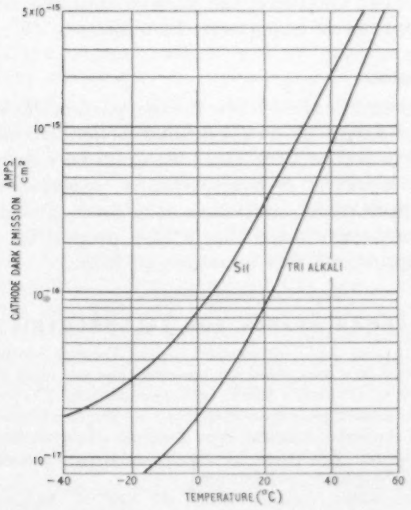


Fig. 15. *Variation of dark emission with temperature*

In place of capacitor integration, the electron pulses at the anode of the photomultiplier tube may be counted with a scaling circuit, but the statistical considerations remain unchanged.

In example (c) of the Section in Part 1 on Measurement of Photocurrent, a television flying-spot system was discussed, in which a cathode current of 0·1 μA was produced, and this, with a bandwidth of 1 Mc/s, gave a signal-to-noise (power) ratio of 1,000. The 15-μV noise in this case was mostly generated by the 10^4-Ω resistor, while the peak-white signal was $\frac{1}{2}$ mV.

With a photomultiplier tube, operated with a gain of 2×10^4 to give a peak anode current of 1 mA from a cathode current of 0·05 μA, the peak white signal would be 10 V in an anode load of 10 kΩ and the r.m.s. noise voltage would be:

$$(1{\cdot}6 \times 10^{-19} \times 2 \times 10^4 \times 10^6)^{\frac{1}{2}}\ \ (4 \times 10^8 \times 10^4 \times 10^{-7} + 0{\cdot}1)^{\frac{1}{2}} = 0{\cdot}03 \text{ V}$$

giving a signal-to-noise (power) ratio of 4×10^4. At zero light level, the noise due to the anode load is 15 μV, while the dark current of about 6×10^{-11} A gives an r.m.s. noise voltage of 1·4 mV. The great improvement due to the use of a photomultiplier tube is shown below:

	Peak White			White Signal / White Noise	White Signal / Black Noise
	Signal	Noise	Black Noise		
PM tube..	10 V	0·08 V	0·0014 V	125	7,000
Photocell	0·5 mV	0·015 mV	0·015 mV	30	30

An example of a statistical problem which would not arise with a photocell, but is important for one widespread application of photomultipliers, is the question of energy resolution in scintillation counters.

These detectors of nuclear radiation consist essentially of a single crystal of a suitable phosphor, such as thallium-activated sodium iodide, optically coupled as closely as possible to the end window of a photomultiplier tube [Fig. 16(a)]. A charged particle which dissipates an energy E electron-volts in the phosphor gives rise to fluorescence in which a total of rE photons are produced, emitted over an exponential decay of time constant τ_1. A proportion R of the light enters the tube photocathode, to give $\eta(RrE) = N$, photoelectrons, where η is the effective quantum efficiency for the fluorescent wavelengths. Over a number of particles, each of which gives the same energy E in the phosphor,

there will be a statistical fluctuation in the observed value of N, the r.m.s. deviation from the mean value being $N^{\frac{1}{2}}$. The mean output voltage, as discussed in the Section above (Photomultiplier-Tube Characteristics), at the anode is $fNGeK_2/C$, where K_2 is the proportion of the maximum value determined by the ratio RC/τ_1 [Fig. 14(c)], so the effective value of the number of photoelectrons involved is K_2fN. The statistical fluctuation due to the independence of production of individual photoelectrons would then result in a pulse-amplitude distribution as in Fig. 16(b), with a width at half amplitude of $\Delta = 2{\cdot}36\ (fNK_2)^{-\frac{1}{2}}\ a \times 100\%$† where a is the enhancement factor discussed in the earlier Section (Noise in Photomultiplier Tubes). An interesting point now emerges. For a typical sodium-iodide crystal on a 2-in. photomultiplier tube, the value of Δ for the photopeak given by a gamma ray of energy 661,000 eV, (Cs^{137}), is $7\frac{1}{2}\%$, but the value of K_2fN deduced from measurement of the voltage pulses, knowing the photomultiplier-tube gain G and the output capacitance C should lead to a Δ of about 4%. This discrepancy suggests that either R, or r, is varying quite widely from scintillation to scintillation, and since tests have shown that Δ is effectively independent of crystal size, which means that R, the average light collection efficiency, is not varying appreciably, one must conclude that the efficiency of conversion of particle energy into light r is not constant from particle to particle. A possible explanation of this variability may be found in the fact that a fast particle transfers part of its energy to medium-velocity electrons (delta rays), which excite the phosphor less efficiently than does the very high velocity primary, and the proportion of energy thus transferred will be subject to statistical variation.

Observation over a range of gamma ray energies with photomultiplier tubes having differing values of photosensitivity, $P\ \mu$A/L, show that $\Delta = 100(90{,}000/EP + 3{\cdot}3/E^{\frac{1}{2}})^{\frac{1}{2}}$, from which it will be seen that the values of P above 40 μA/L have very little effect on Δ for $E > 100{,}000$ eV, but for $E < 20{,}000$ eV the photoelectron statistics become of greater importance than the internal spreading factor, $3{\cdot}3/E^{\frac{1}{2}}$, and it is then desirable to choose a photomultiplier tube with as high a value of photosensitivity as possible, preferably >70 μA/L. In this latter case the tube must also have a low dark current to avoid an unduly high background due to thermionic electrons.

An extreme case in which the background is of importance is the scintillation counting of low energy β emitters in a liquid phosphor, tritium having such low energy that only 60–70% of the disintegrations give rise to a photo-electron from the cathode and single electrons due to scintillations are of course indistinguishable from the thermionic electrons. For such applications, tubes with specially processed SbCs cathodes having minimum thermionic emission (E.M.I. S cathodes) are produced.

Applications of Photomultiplier Tubes

From the final Section of the first part of this paper, it will be clear that one important application of photomultiplier tubes is in the measurement of light levels so

† A Gaussian curve of standard deviation σ had a width at half amplitude of $2{\cdot}36\sigma$.

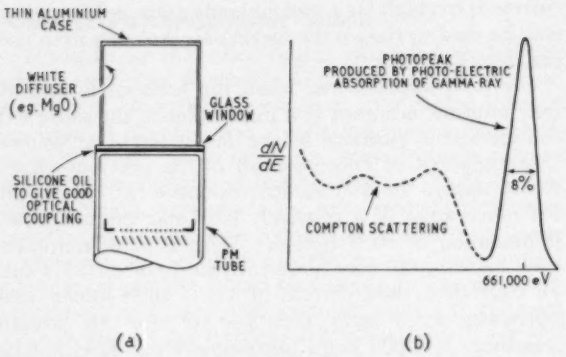


Fig. 16. *Shows the construction of a nuclear radiation detector which is NaI-Tl crystal (in an aluminium can) mounted on a p.m. tube (a); pulse amplitude distribution from Cs^{137} γ-rays on NaI-Tl is shown in (b)*

low that Johnson noise in the load resistor precludes the use of a photocell, corresponding to a cathode current of the order of $C/20t$, or less, where C is the capacitance and t the response time, of the measuring circuit. For a photosensitivity of 60 μA/L and capacitance of 25 pF, this corresponds to a light level of $2 \times 10^{-8}/t$ lumens.

When the photocell is replaced by a photomultiplier tube, the comparable limit is smaller by a factor $1/G^2$, but the effective value under these conditions is set either by dark current, as in Table 5, or statistically at $4e/t$ A, which gives a signal-to-noise (current) ratio of 2 : 1, and corresponds to $10^{-14}/t$ lumens for 60 μA/L. Dark current sets the limit in many astronomical, spectrophotometric and fluorimeter applications when the very lowest light levels are to be measured, with values of t greater than a second. The statistical limit is effective in many spectrophotometric instruments using a.c. measuring circuits and, of course, in the transducer applications of which flying-spot television pick-up is an example.

The other main class of application calls into play the ability of a photomultiplier tube to respond to short light flashes without time degradation and is exemplified by scintillation counters, and some other transducer applications. Of course, the dark current and statistical limitations are of importance here also.

Conclusion

It has been the aim of these the two parts of this paper to present a survey of photoemissive devices mainly from the statistical viewpoint, since the granular nature of the physical world is of overwhelming importance in so many applications, and experience over a number of years has demonstrated that a high proportion of users fail to appreciate this fundamental limit.

GERMAN AGENCY FOR TELEMECHANICS

Telemechanics Ltd. (Brokenford Lane, Totton, Southampton) have recently been appointed sole agents for the sale of all electronic equipments of Deutscher Innen und Aussenhandel Elektrotechnik, Berlin, German Democratic Republic, in the territories United Kingdom, Australia, Canada, New Zealand, Malayan Federation, Central African Federation of North and South Rhodesia and Nyasaland, Singapore and Portugal.

The equipments broadly cover the field of frequency and time measurement, a.c.-signal generation, and C, R and L measurements.

Phase-Sensitive Wave Analysis

USING ANALOGUE-COMPUTER TECHNIQUES

By O. J. Haga,* B.Sc., and D. Midgley†, Ph.D.

SUMMARY. *A method of phase analysis is described in which the input waveform is applied to one input of a multiplier. The other input takes in turn reference sine and cosine waves generated locally at the harmonic frequencies. To maintain phase-coherence, operation is interrupted in alternate fundamental periods, allowing initial conditions to be reset in the reference-wave generator.*

Results are given for an experimental circuit with a wattmeter performing the function of multiplication.

Wave analysers usually depend for their action upon some form of selective circuit, which introduces a rapid change of phase as the selected frequency is approached. For this and other reasons, it is hard to find a wave analyser which performs a complete Fourier analysis of its input[3,5]. Only magnitudes of the harmonics are found and this is sufficient for a number of applications, such as vibration analysis where coherent phase relations between the frequency components may not in fact exist. The simple wattmeter method is distinguished by the absence of selective circuits, but the information contained in the phase spectrum is again lost.

The phase spectrum is equal in importance to the amplitude spectrum in determining the shape of the original wave. For instance, the waveform of the current in an iron-cored coil with a sinusoidal e.m.f. has a characteristic shape which disappears if the phases of third and fifth harmonics are disturbed. A waveform synthesizer certainly cannot dispense with phase. The present work began as an attempt to invert a useful waveform synthesizer[1], which had on occasions been pressed into service for the reverse process of analysis. This consisted in the skilful setting of potentiometers controlling sine and cosine components of each harmonic until the synthesized waveform approximated a picture of the waveform to be analysed. The process is tedious to say the least. Whereas a synthesizer must have many coherent oscillators acting simultaneously at fixed frequencies, an analyser need only generate one frequency at a time. Both processes can use a ringing circuit with zero damping. When analogue-computer amplifiers are available, a natural choice among such circuits is a two-integrator version based upon the oscillator of E. F. Good[2]. This has both sine and cosine output points and such a feature is necessary to the proposed method of analysis, which is indicated in Fig. 1.

Some feature of the waveform under examination is chosen to initiate a blanking wave. The latter decides the start of the sine and cosine waves and alters the damping so as to extinguish each burst of oscillation, after an integral number of cycles with zero damping. The undamped parts of these waveforms are then compared in a multiplier with the input waveform.

The result is in fact the application of an analogue computer to perform the mathematical process of

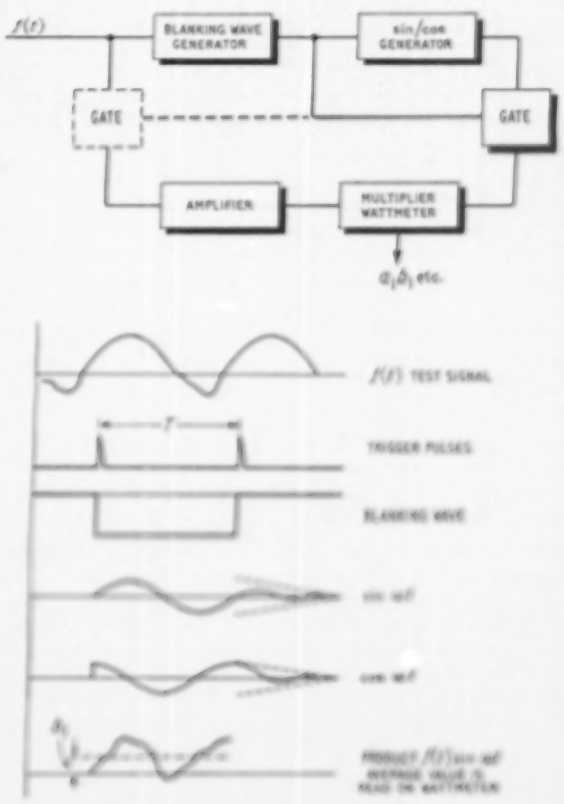


Fig. 1. *Block diagram and waveforms.*

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† Reader in Electrical Engineering at the Bradford Institute of Technology.

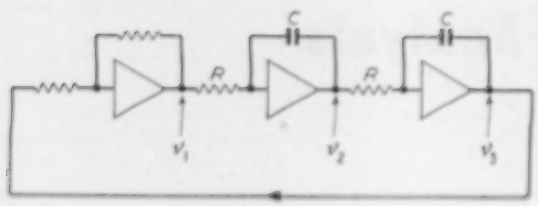

Fig. 2. A two-integrator loop

representing a periodic waveform $f(\omega t)$ as a Fourier Series in the a_n, b_n form:

$$f(\omega t) = \sum_{n=1}^{\infty} [a_n \cos(n\omega t) + b_n \sin(n\omega t)] + a_0$$

$$a_n = \frac{1}{\pi} \int_{-\pi}^{\pi} f(\omega t) \cos(n\omega t)\, d(\omega t)$$

$$b_n = \frac{1}{\pi} \int_{-\pi}^{\pi} f(\omega t) \sin(n\omega t)\, d(\omega t)$$

for $f(\omega t)$ is multiplied in turn by locally-generated $\frac{\sin}{\cos}(n\omega t)$ quantities and the average of each product is recorded as a term in the series.

Circuit Description

The input waveform is applied to a Schmitt trigger circuit V_1 (Fig. 4), the differentiated output from which operates a bi-stable multivibrator V_3. The result is a square blanking wave, one-half period of which covers one period of the input waveform. The intention is that computation shall occupy half the time, leaving intervals free for the decay of oscillation in the ringing circuit and for the re-setting of initial conditions.

The ringing circuit is essentially a loop containing two integrators and one inverter (Fig. 2). Finite gain in the amplifiers and impurity in the components prevent such a single loop from possessing zero damping in practice. Unlike many oscillatory feedback systems, the loop gain controls the frequency of oscillation rather than the damping. However, an additional resistive feedback connection from the output of the first integrator to the summing junction of the inverter introduces negative damping (Fig. 3 and Appendix). To simplify control of the damping, particularly at high frequencies, the capacitor of the first integrator is deliberately degraded by the inclusion of resistance. Those stray capacitances which tend to give uncontrolled instability at high frequencies are thus defeated by the existence of excess initial damping, which is readily compensated by the controlled negative damping of the inner loop. The second integrator is not treated in this way, since it is required to maintain an exact phase relation between the sine and cosine outputs.

The initial-condition circuits re-charge the two capacitors in the ringing circuit during the reset period to pre-determined initial voltages, which determine the amplitude and phase of the next oscillation. Theoretically-perfect integrators would require only one capacitor to be charged initially for exact sine and cosine outputs, and the initial voltage would control only the amplitude of the oscillation. It is preferable in practice to provide the two initial voltages allowing fine adjustment of the phase of the oscillation to be made, so that sine and cosine outputs can be set accurately at every frequency despite the presence of imperfections such as finite gain and significant stray capacitances. A number of alternative initial-condition circuits exist[4]. A low internal resistance in the charging circuit is desired to effect full damping in as short a time as possible. Diodes would offer a smaller resistance, but they would demand more power than triodes from the source of the blanking wave and they would introduce a harmful shunt resistance across the capacitor of the second integrator during the computation period. The head-to-tail triode circuit depends upon both grids being driven well below cut-off during computation, by the negative-half of the blanking wave, which is delivered through an amplifier and clamping circuit.

A wattmeter acts as both multiplier and output indicator in the experimental arrangement. It requires calibration against frequency and a power amplifier to drive the current coils. Error is introduced especially on the low-voltage ranges at high frequency by the mutual inductance between the voltage and current coils.

Experimental Results

The procedure for performing an analysis is as follows:

1. The given voltage is applied to one of the inputs to the multiplier, to the blanking-wave generator and to synchronize the time base of an oscilloscope.
2. The blanking waveform is examined on the oscilloscope, and the voltage at the grid of the Schmitt trigger circuit is adjusted until the start of the blanking wave coincides with some chosen point of the input waveform. This choice must remain fixed throughout the analysis.
3. The sine output of the ringing circuits is displayed on the oscilloscope and the frequency is adjusted until n complete cycles are seen. Damping is adjusted so that each cycle has the same peak value, preferably while the cosine output is displayed.
4. The initial-condition controls set the amplitudes of the sine and cosine waves to a standard value. They also are adjusted to set the exact phase of, preferably, the sine output; the other output should then be an exact cosine.
5. The sine waveform is applied to the second input of the multiplier or wattmeter, the steady output of which is the required term b_n in the analysis. Switching to the cosine waveform gives a_n.

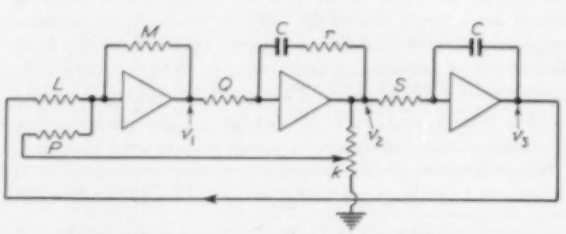

Fig. 3. A two-integrator loop with damping control

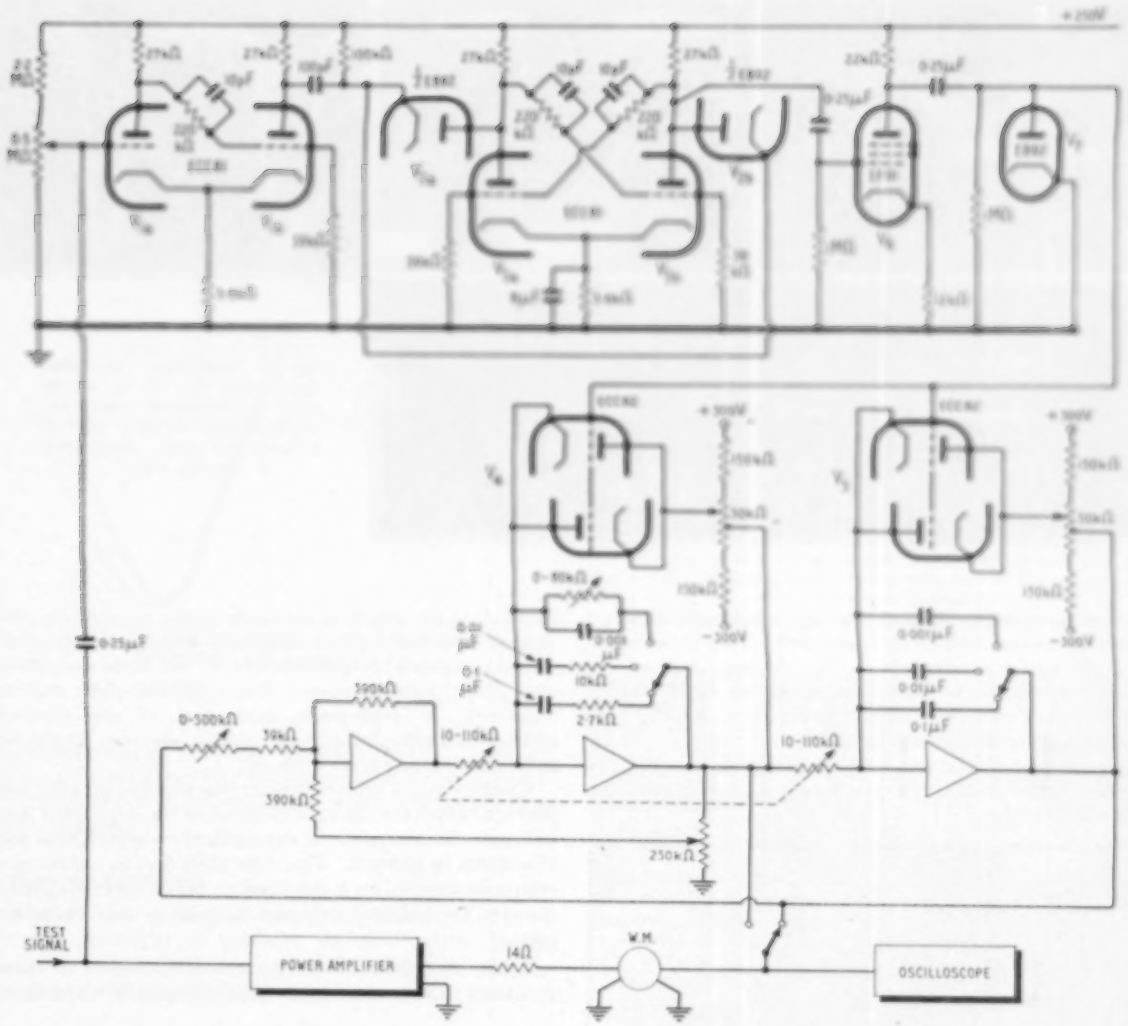

Fig. 4. *Circuit diagram showing blanking-wave generator ($V_1 - V_3$) and initial-condition circuits (V_4 and V_5)*

Steps 3, 4, and 5 are repeated for other integral values of n.

Oscillograms of the blanking wave, sine and cosine outputs appear in Fig. 5. Damping is so effective that it obliterates the unwanted parts and it is not necessary to include gates between the ringing circuits and the multiplier. The initial-condition circuits will perform more than 1,000 resets per second, which is a higher speed than is likely to be required. The ringing circuit requires careful adjustment at the upper end of its high-frequency range 1–10 kc/s, every change in frequency then requires a re-adjustment of the damping.

An initial trial designed to test the phase facility of the analyser consisted in the analysis of a waveform having fixed amplitudes of fundamental and third harmonic, but continuously-variable phase. Two such characteristic waveforms representing the extreme variation of phase are shown in Fig. 6. Advancing the phase of the third harmonic in nominal 45° steps through 135° gave the following results:

	a_n	b_n	c_n	$\tan \phi_n$	ϕ_n	Increment in ϕ_3
Fundamental	50	−25	56	+2·0	244°	
Third harmonic	−34	−74	81	−0·46	155°	
	31	−77	83	0·40	202°	47°
	77	−32	83	2·41	247°	45°
	70	37	79	−1·89	298°	51°
	26	74	79	−0·35	341°	43°
	−45	68	82	0·66	033°	52°

There is no ambiguity in determining the angle from its tangent, since the quadrant is fixed by the signs of a_n and b_n. If $a_n \cos(n\omega t) + b_n \sin(n\omega t) = c_n \sin(n\omega t - \phi_n)$, it follows that $a_n = -c_n \sin \phi_n$; $b_n = c_n \cos \phi_n$; $\tan \phi_n = -a_n/b_n$; and $c_n = \sqrt{a_n^2 + b_n^2}$. The results demonstrate the ability of the system in dealing with phase measurement in every quadrant.

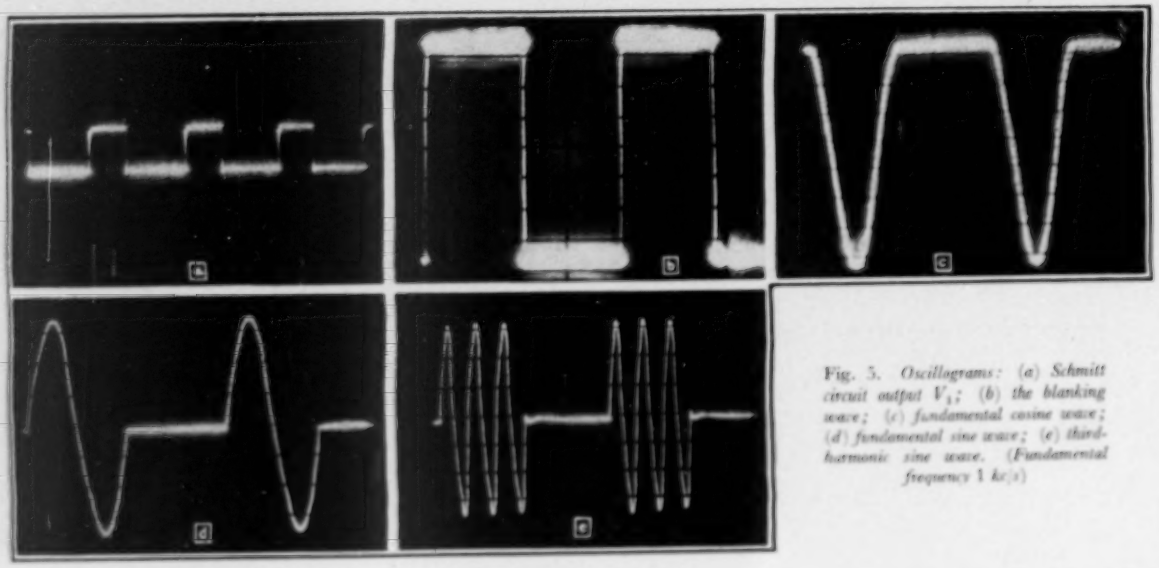

Fig. 5. *Oscillograms: (a) Schmitt circuit output V_1; (b) the blanking wave; (c) fundamental cosine wave; (d) fundamental sine wave; (e) third-harmonic sine wave. (Fundamental frequency 1 kc/s)*

The amplitudes found for the nominally-constant third harmonic deviate by up to 2·5% from their mean, and the phase increments by 5°. At least half of this error is attributed to the source of the test signal, which generates an imperfect fundamental wave (Fig. 6), and lacks accurate phase-calibration.

A more complete analysis of a waveform (Fig. 7) derived from a half-wave rectified 50-c/s source gave the following results:

n	a_n	b_n	$\tan \phi_n$	ϕ_n	Angle referred to fundamental	Amplitude referred to fundamental
1	123	−142	0·87	221°	0°	1
2	29	−69	0·42	203°	−18°	0·40
3	01	−05	0·2	191°	−30°	0·027
4	−04	−15	0·27	165°	−56°	0·085
5	. 06	06	1·00	045°	−176°	0·042
6	−08	06	1·33	053°	−168°	0·045

The relative amplitudes and phases of the three most significant components (1st, 2nd and 4th) are illustrated in Fig. 7 and a Fourier approximation curve derived from them is compared with the original waveform.

Discussion

A method of complete Fourier analysis has been demonstrated, using very little equipment additional to that which is normally available at a small analogue computer. When measuring harmonics, it is possible to achieve an accuracy of at least 2% of the fundamental amplitude and 2° of the fundamental phase angle.

Although the method may be regarded as an extension of the wattmeter technique to include phase measurement, virtually any other type of multiplier would be preferable, unless the wattmeter could be compensated against errors at high frequencies.

The absence of selective circuits and heterodyne oscillators is a merit of the present method, the setting-up procedure for which is certainly easier to perform than that of some heterodyne analysers. Manipulation of two initial-condition potentiometers to set both amplitude and phase (item 4, above) is less difficult than may be supposed; a push-push movement of the controls changes amplitude without phase, whereas push-pull movement controls only phase.

Quenching of the reference waveforms in alternate periods halves the mean indication of the multiplier and prevents the extension of the method to waveforms and vibrations in general. The only objection to continuous reference waveforms is the need to devise more elaborate circuits for holding constant amplitude and coherent phase.

Thus, the present arrangement is restricted to those functions which are strictly periodic and so capable of

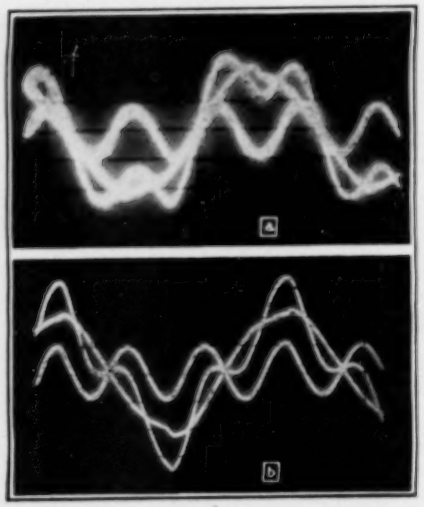

Fig. 6. *A test waveform and its harmonics (a) peaks of fundamental and a third harmonic opposing, and (b) aiding*

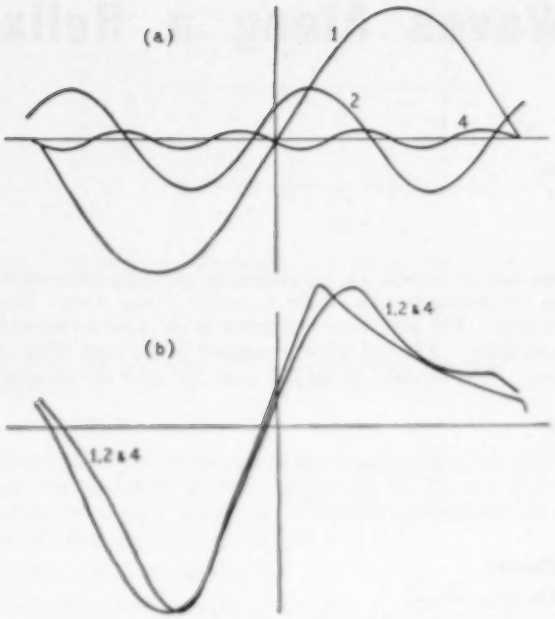

Fig. 7. (a) The three principal components of a test waveform and (b) their sum compared with the original wave

representation by a Fourier Series as distinct from the Fourier Transform. The principal merit is the inclusion of phase-spectrum information, which otherwise is so often neglected.

Acknowledgements

The authors wish to thank Professor E. G. Cullwick for the provision of facilities for this work in the University of St. Andrews, and Mr. W. R. McEwan for assistance with photography.

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APPENDIX

Analysis of the Two-Integrator Loop

Let v_1 v_2 v_3 be the output voltages of the three amplifiers. A fraction kv_3 is fed-back for reducing damping and a fixed resistor r is added in series with the capacitor of the first integrator to bring the initial damping, when k is zero, over to the positive side at all frequencies.

$$v_1 = -\frac{M}{L} v_3 - \frac{Mkv_3}{P} \quad \ldots \quad (1)$$

$$v_2 = -\frac{1}{pCQ} v_1 - \frac{r}{Q} v_1 \quad \ldots \quad (2)$$

$$v_3 = -\frac{1}{pCS} v_2 \quad \ldots \quad (3)$$

Substituting Equ. (3) in Equ. (1) and then Equ. (2) in Equ. (1), v_2 and v_3 are eliminated . . .

$$v_1 = \left(\frac{M}{pCLS} - \frac{Mk}{P}\right) v_2 = \left(\frac{M}{pCLS} - \frac{Mk}{P}\right)\left(-\frac{1}{pCQ} - \frac{r}{Q}\right) v_1$$

$$= \left(-\frac{M}{p^2C^2LSQ} + \frac{Mk}{pPCQ} - \frac{rM}{pCLSQ} + \frac{rMk}{PQ}\right) v_1$$

$$p^2C^2LSv_1 - p^2C^2LS\frac{rMk}{PQ} v_1 + pC\frac{rM}{Q} v_1 - pCLS\frac{Mk}{PQ} v_1 + \frac{M}{Q} v_1 = 0$$

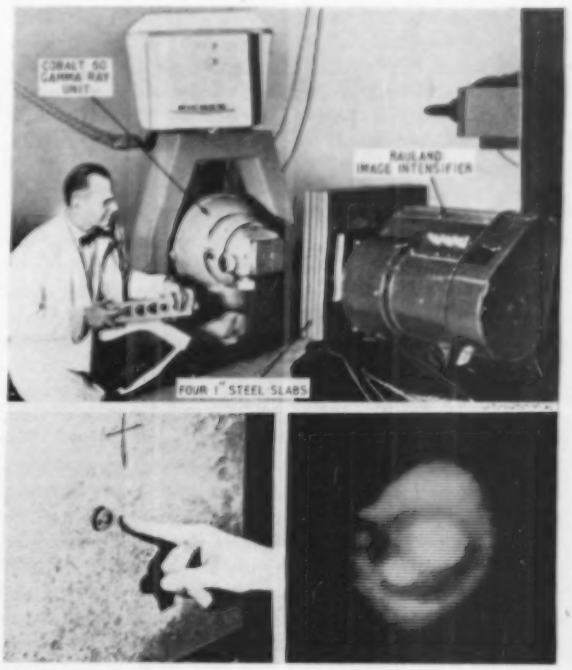
This is the differential equation for damped harmonic motion. The third term shows the damping due to r and the fourth term shows how it may be cancelled by setting $k = rP/LS$. The second term reveals an effect upon frequency, which disappears when either r or k is zero, and reveals the desirability of keeping r as small as possible.

NEW GAMMA-RAY METAL FLAW DETECTOR

The Rauland Corporation and Picker X-ray Corporation, U.S.A., have developed a gamma-ray image-intensifier system which is designed to detect minute flaws in steel up to 4 in. thick and other dense objects and to display images of the flaws on a television screen.

Used in conjunction with the Picker Orthicon camera and 'Cyclops' Cobalt 60 unit, the Rauland intensifier tube will permit continuous inspection of dense materials which previously could be inspected only by gamma-ray photographs, and of other materials which could not be inspected without destroying them.

In operation, gamma rays from the Cobalt 60 unit are directed



through the material being inspected to a thin metal plate at one end of the Rauland tube. Here some of the energy is transferred to electrons of low enough intensity to produce a fluoroscopic image which is intensified by the tube to produce a picture at the other end about 1,000 times brighter, sufficient to provide a clear picture on a remote television screen or to be viewed directly through an optical system.

The illustrations show the complete equipment, a 1-in. diameter steel disc $\frac{1}{8}$ in. thick with holes in it $\frac{1}{8}$ in. and $\frac{1}{4}$-in. diameter which is used to simulate a flaw in four 1-in. thick steel plates, and a photograph of the television display of this flaw.

Propagation of E.M. Waves Along a Helix

EFFECTS OF AN INNER CONDUCTOR

By R. Dingeldey,* Dipl. Ing. and A. Cunliffe,* B.Sc., Ph.D.

SUMMARY. *The introduction of an inner conductor along the axis of a helix for the purpose of providing electrostatic focusing in a travelling-wave tube is shown to form a system which can propagate a fast wave as well as a slow wave. The properties of these waves are investigated theoretically and experimentally. The slow wave corresponds to the wave propagated along the helix alone, but modified by the presence of the inner conductor. The fast wave corresponds to the wave along a single straight wire, but modified slightly by the presence of the helix. The existence of the fast wave can affect the operation of a travelling-wave tube.*

Long electron beams are usually focused magnetically but electrostatic methods of focusing have also been considered. One electrostatic method has become known as 'Harris-flow'[1]. In it the electrons travel along helical paths between two coaxial cylindrical electrodes, the inner one of which is positive with respect to the outer one. The resultant radial electric field is formed from the potential difference applied between the two electrodes and the space charge of the electron beam. The radial electric field gives rise to a force which balances the centrifugal force which is associated with the electrons because of the circumferential component of their velocity. The space occupied by the electron beam is then a hollow cylinder coaxial with the system, the inner radius of the cylinder being larger than the radius of the inner electrode, and the outer radius of the cylinder being smaller than the inner radius of the outer electrode. This method of focusing an electron beam without an axial magnetic field has been used with the 'Helitron Oscillator'[2], and the 'Spiratron'[3].

In the case of the Spiratron the outer electrode takes the form of a helix which also serves as the slow-wave structure as in a normal travelling-wave tube. The inner electrode is a straight wire introduced along the axis of the helix as in Fig. 1. The presence of this wire means that the wave-propagation characteristics of the system are different from those of the usual travelling-wave tube system which has only the helix. It is shown in this article that two types of waves can be propagated along the modified system. One of these waves is a slow wave like the wave propagated along the helix alone, modified only slightly by the presence of the inner wire. The other wave is a fast wave like the wave propagated along a single straight wire, modified only slightly by the presence of the helix.

When such a system is used in a travelling-wave tube, the slow wave will interact with the electron beam but the fast wave will not. The existence of the fast wave can, however, affect appreciably the operation of the travelling-wave tube. A method of preventing the excitation or propagation of the fast wave is required.

* *Physics Department, University of Hull.*

Theory

The Slow Wave

The theory developed by Pierce[4] for the propagation of slow waves along a helix requires modification when an inner conductor is present. Choosing cylindrical co-ordinates r, ϕ, and z, with the z-axis along the axis of the helix, the components E_z and H_z of the electric and magnetic fields will be subject to the equations

$$\left.\begin{aligned}\frac{\partial^2 E_z}{\partial r^2}+\frac{1}{r}\frac{\partial E_z}{\partial r}+\left(\frac{\omega^2}{c^2}-\beta^2\right)E_z=0\\ \frac{\partial^2 H_z}{\partial r^2}+\frac{1}{r}\frac{\partial H_z}{\partial r}+\left(\frac{\omega^2}{c^2}-\beta^2\right)H_z=0\end{aligned}\right\} \quad .. \quad (1)$$

where ω is the angular frequency of the oscillations, c is the velocity of plane waves in vacuo, and $\beta = \omega/v$ where v is the phase velocity of the slow wave.

Fig. 1. *Helix with an inner conductor*

When there is no central conducting wire, modified Bessel functions of the second kind are not involved in the solution of these equations for the space lying inside the helix, but this is no longer the case when the inner conductor is present. The subscripts i and o will be used to denote the space between the inner conductor and the helix, and the space outside the helix respectively. The z components of the electric and magnetic fields can then be written in the form

$$\left.\begin{aligned}E_{zi}&=A_1I_o(\gamma r)+B_1K_o(\gamma r)\\ H_{zi}&=A_2I_o(\gamma r)+B_2K_o(\gamma r)\\ E_{zo}&=B_3K_o(\gamma r)\\ H_{zo}&=B_4K_o(\gamma r)\end{aligned}\right\} \quad .. \quad .. \quad .. \quad (2)$$

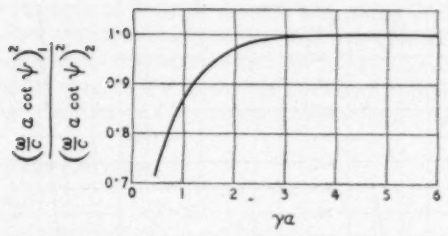


Fig. 2. *Ratio of the parameters* $[(\omega/c)\ a\ \cot\psi]^2$ *of a helix with an inner conductor (subscript 1) and one without an inner conductor (subscript 2)*

where the As and Bs are constants times $e^{j(\omega t-\beta z)}$ and

$$\gamma^2 = \beta^2[1-(v/c)^2] = \beta^2 - (\omega/c)^2 \quad .. \quad .. \quad (3)$$

The radial and circumferential components of the fields can be calculated in terms of A_1, A_2, B_1, B_2, and B_3 by using Maxwell's equations. The field components must satisfy the boundary conditions

$E_{zi} = 0 = E_{\phi i}$ at $r = b$;

$E_{zi} \sin\psi + E_{\phi i} \cos\psi = 0$,

$E_{zi} = E_{zo}$, $E_{\phi i} = E_{\phi o}$, and

$H_{zi} \sin\psi + H_{\phi i} \cos\psi = H_{zo} \sin\psi + H_{\phi o} \cos\psi$ at $r = a$.

The boundary conditions for the electric-field components fix the relative values of the As and Bs, so that expressions for all the field components can be obtained which involve A_1 only. These expressions are listed in Equs (A1) of the Appendix. The power transported along the z-direction can be obtained in terms of A_1 from the radial and circumferential-field components, and this is given in Equ. (A2). The boundary condition for the magnetic-field components yields a transcendental equation from which γ can be calculated. This equation is

$$\gamma^2 = \frac{a_4^2}{a_3^2}\left[\frac{a_2/a_4 - b_2/b_4}{a_1/a_3 - b_1/b_3}\right]\left(\frac{\omega}{c}\cot\psi\right)^2 \quad .. \quad .. \quad (4)$$

where the as and bs, which involve γ, are defined in Equs (A3) of the Appendix.

The quantities which are of importance from the point of view of using the system as the delay line of a travelling-wave tube are γ, and the electric- and magnetic-field components in the region of the electron beam for a given power transported along the system. It is therefore of interest to compare these quantities for a helix with an inner conductor with the corresponding quantities for a helix of similar pitch and radius but which has no inner conductor. If b_1/b_3, b_2/b_4 and b are put equal to zero in Equs (A1), (A2) and (4), the equations reduce to those appropriate to a helix without an inner conductor, as listed by Pierce.

It is difficult to calculate values of γ from Equ. (4), and so the following procedure is used to find the effect of the inner conductor. Two systems having equal values of γa are considered: one system consists of a helix with an inner conductor having a radius equal to one-tenth of the radius of the helix, and the other system consists of a helix alone. Equ. (4), and the corresponding equation obtained by putting b_1/b_3 and b_2/b_4 equal to zero in Equ. (4), are used to find the ratio of the parameters $[(\omega/c)\ a\ \cot\psi]^2$ for these two systems for different values of γa. This ratio is plotted against γa in Fig. 2. From this figure it can be seen that for $\gamma a > 3$, the parameters $[(\omega/c)\ a\cot\psi]^2$, which contain the dimensions and the frequency, do not differ appreciably. This means that for $\gamma a > 3$, two helices with equal as and ψs, but one with an inner conductor and the other one without, effectively have equal γs.

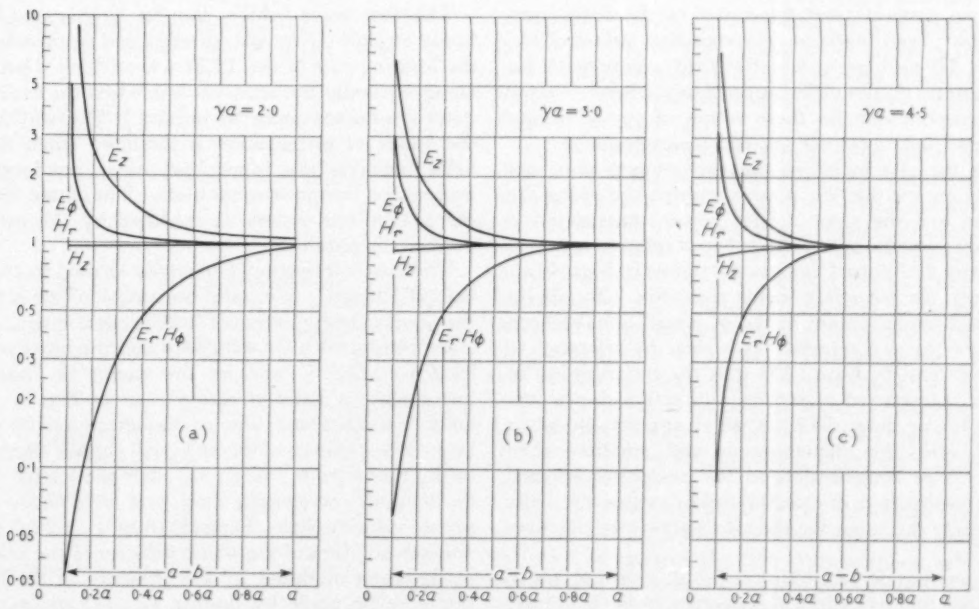


Fig. 3. *Ratios of the field components of a helix without an inner conductor and those of a helix with an inner conductor*

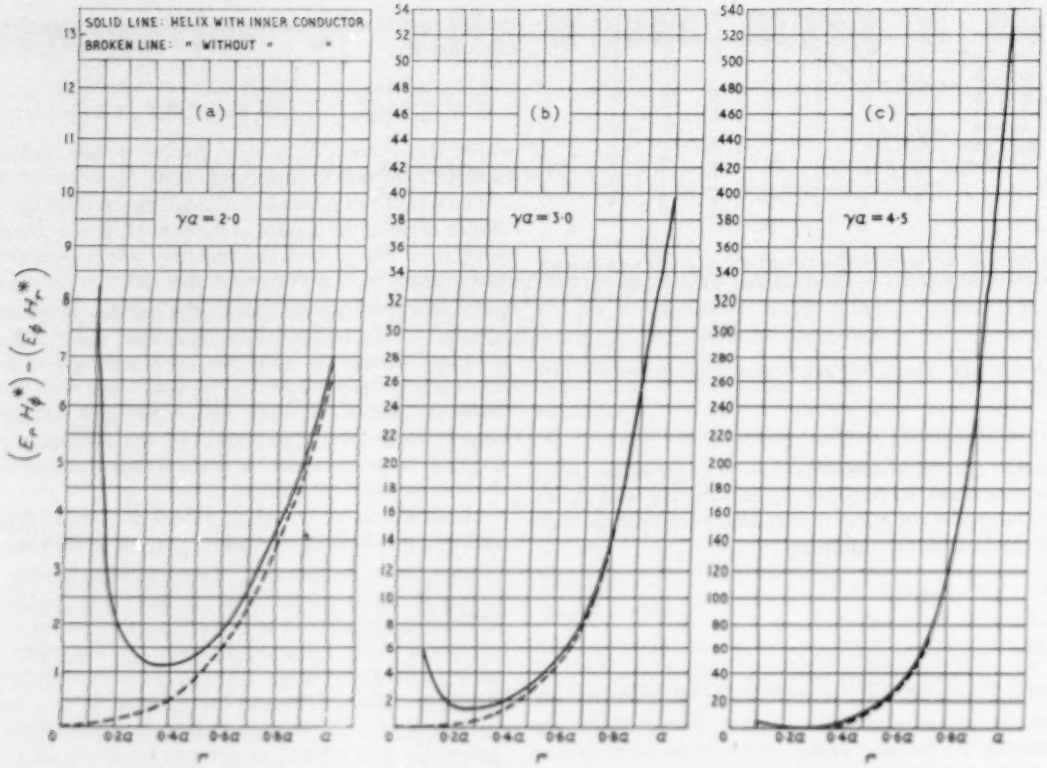


Fig. 4. $E_r H_\phi^* - E_\phi H_r^*$ *plotted against radius* r *for three values of* γa*; solid line, helix with inner conductor; dashed line, helix without inner conductor*

The ratio of the field components for the helix without the inner conductor and the helix with the inner are plotted against the radius in Figs. 3 (a), (b) and (c), for the space inside the helix. To obtain each of these curves, equal values of A_1 and γa have been assumed for the two systems, and the ratios of the field components have been taken at corresponding values of r/a. For $\gamma a = 3{\cdot}0$ and $\gamma a = 4{\cdot}5$, the field components for the two systems do not differ appreciably when $r > 0{\cdot}6a$. It has been seen that for these values of γa the design parameters $[(\omega/c)\, a \cot \psi]^2$ are very nearly equal.

Putting the A_1s equal in the two systems does not necessarily ensure that the powers transported along the z-direction are the same. The power transported is determined from $E_r H_\phi^* - E_\phi H_r^*$ and relative values of this quantity are plotted against the radius in Figs. 4 (a), (b) and (c), for the space inside the helix. To obtain these values, equal values of A_1, γ, a and ω have been assumed for the two systems. It is clear by reference to these curves and to Equ. (A2) that the contribution to the power transported along the z-direction due to the fields inside the helix does not differ appreciably when $\gamma a = 3{\cdot}0$, and the difference is still smaller when $\gamma a = 4{\cdot}5$. The contribution to the power transported along the z-direction due to the fields outside the helix is very nearly the same for the two systems for all three values of γa.

It follows that if $\gamma a > 3$, $b = 0{\cdot}1\ a$ and the mean radius of the hollow cylindrical electron beam is greater than $0{\cdot}6a$, the presence of the inner conductor should not affect appreciably the interaction between the electron beam and the r.f. field in a travelling-wave tube as far as the slow wave is concerned.

The Fast Wave

The fast wave which can be propagated along a single straight wire can be regarded approximately as the limiting case of the TEM wave which is propagated along a coaxial transmission line when the radius of the outer conductor tends to infinity. The way in which this mode of propagation is modified when there is a helix between the inner and outer conductors is of interest for the present problem. The phase velocity of the coaxial-line system as modified by the presence of the helix is required.

The lowest frequency of a cavity formed by closing the ends of a length l of coaxial transmission line is $f = c/2\ l$, the cavity being one-half wavelength long. Let the outer conductor have a radius d and the inner conductor have a radius b, and let the cavity be modified by introducing a helix of radius a as in Fig. 5. Let the helix be formed from wire of diameter e and let the pitch be p. The presence of the helix will change the frequency of the mode from f to f_p, say, but the cavity must still be one-half wavelength long, and this means that the phase velocity has changed from $c = 2\,l f$ to $2\,l f_p$. A determination of the phase velocity of the coaxial-line mode when modified by the presence of the helix can therefore be made by finding f_p. f_p can be estimated by using perturbation theory. For $l = 46{\cdot}5$ mm,

$b = 0{\cdot}3$ mm, $a = 3$ mm, $e = 0{\cdot}25$ mm and $p = 1$ mm, the theory indicates that f_p and f differ by considerably less than one per cent when d exceeds $3a$. It is concluded that the presence of the helix does not affect appreciably the phase velocity of the transmission-line type mode.

Experimental Work

Apparatus

The existence of slow and fast waves on a helix with an inner cylindrical conductor was demonstrated by means of the apparatus shown in Fig. 6. The reflex klystron oscillator O operating at a free space wavelength of about 9 cm was coupled to the helix H via an attenuator D, a slotted line M and a short length of counter-wound helix K. A test probe S, which could be moved either along the slotted section or the helix, fed energy to a crystal detector and then to a galvanometer G. The probe system could therefore be used to indicate the standing waves in M, but its main purpose was to detect the radial electric field on the helix arrangement. Details of the helix arrangement are shown in Fig. 7. The helix was wound from copper wire of diameter 0·5 mm and had a diameter of 6 mm and a pitch of 1 mm. The helix was placed for support inside a glass tube, the thickness of the glass being approximately 0·75 mm, and a copper wire of diameter 0·6 mm could be placed along the axis of the tube. A second piece of counter-wound helix at the opposite end of the helix from where the oscillations were introduced, ensured that standing waves were produced in the helix system. The frequency of the source was monitored by the cavity R and microammeter.

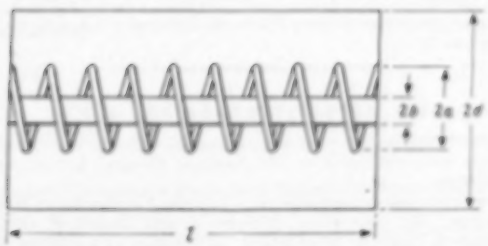


Fig. 5. *Coaxial cavity resonator perturbed by a helix*

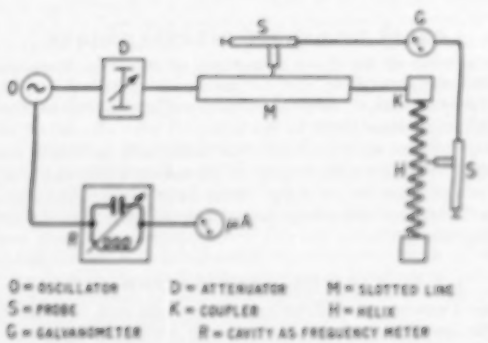


Fig. 6. *Apparatus used to demonstrate the existence of slow and fast waves on a helix with an inner cylindrical conductor*

Fig. 7. *Experimental helix arrangement with an inner conductor*

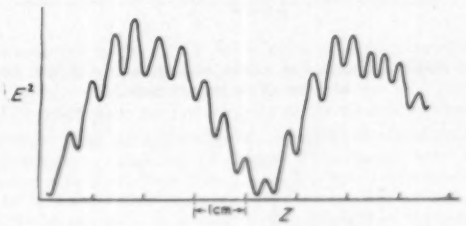


Fig. 8. *Radial electric field of a helix with an inner conductor, at the glass envelope*

Measurements

The galvanometer deflection was measured as a function of the distance of the test probe along the helix, and the results obtained are plotted in Figs. 8, 9 and 10.

Fig. 8 was obtained when the central wire was present, and in this experiment, the tip of the test probe just touched the glass envelope surrounding the helix. Since the fields of both the slow wave and the fast wave fall off with increasing distance from the helix, this graph is for the largest field which can be detected. From the figure it is clear that there is a slow wave of wavelength 0·9 cm superimposed on a larger fast wave of wavelength 9·1 cm.

Fig. 9 was obtained with the central wire present, but the tip of the test probe was withdrawn away from the helix so that it was a few millimetres from the glass envelope surrounding the helix. Only the fast wave now appears. This is in accordance with theory, according to which the field strength associated with the fast wave falls off with radius as $1/r$ whereas the radial electric field associated with the slow wave varies as $K_1(\gamma r)$. The fast wave therefore predominates some distance from the helix.

In the measurements which have been described above, the helix was approximately matched to the transmission line feeding it, the voltage standing-wave ratio measured by means of the slotted section and test probe being 2·00. By altering the frequency of the source from the value 3·227 kMc/s at which this v.s.w.r. was obtained, the matching of the helix to the line could be varied. Thus at frequencies of 3·239 and 3·244 kMc/s, the v.s.w.rs were 2·96 and 3·18 respectively. Patterns very similar to Figs 8 and 9 were obtained at these two frequencies and, from this, it appears that the relative amount of excitation of the two modes cannot easily be varied by altering the matching of the helix to the input transmission line.

Fig. 10 was obtained with the inner conductor removed, but with the conditions otherwise the same as those for obtaining Fig. 8. When there is no inner conductor, the radial electric field shows up only the presence of the slow wave, so the fast wave must be a consequence of the inner conductor.

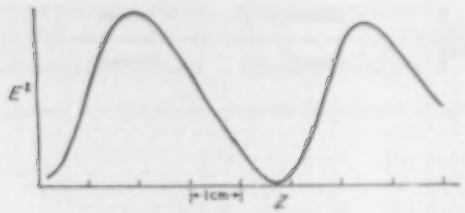


Fig. 9. *Radial electric field of a helix with an inner conductor, measured at 8 mm off the glass envelope*

Fig. 10. *Radial electric field of a helix without an inner conductor*

Conclusions

The introduction of an inner conductor along the axis of a helix for the purpose of producing electrostatic focusing in a travelling-wave tube has been shown to form a system which can propagate a fast wave as well as the slow wave. Under certain conditions, the equations which are valid for the slow wave which is propagated along a helix without an inner conductor may be taken as good approximations for the corresponding equations for the slow wave which is propagated along a helix with an inner conductor. In a travelling-wave tube, the fast wave which can be propagated when there is an inner conductor will not couple to the electron beam, but there can be direct transmission of the input signal to the output via the fast wave. This direct transmission via the fast wave cannot be conveniently prevented by a graphite coating on the glass envelope, and may affect the performance of low-gain high-power travelling-wave tubes. It is desirable to introduce some mechanism into the tube to either suppress the fast wave, or to prevent its being excited.

Acknowledgements

The authors wish to express their thanks to Professor Garlick for his interest in their work, and Mr. N. Taylor for the accurate making of the helix arrangement and other parts of the apparatus. One of the authors (R.D.) is very grateful to the British Council for a scholarship which enabled him to work in the Physics Department of the University of Hull during the session 1959–60, and he also wishes to thank his employers, the Deutsche Bundespost, for the leave granted for this period.

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APPENDIX

The electric- and magnetic-field components for the helix with an inner conductor are as follows.

$$
\left.\begin{aligned}
E_{zi} &= jA_1(\beta/\gamma)\,[r_2 + (b_2/b_4)r_4] \\
E_{\phi i} &= -A_1 \tan\phi \left[\frac{a_1 - (b_1/b_2)a_2}{a_2 - (b_2/b_4)a_4}\right][r_2 - (b_2/b_4)r_4] \\
E_{ri} &= A_1\,[r_1 - (b_1/b_2)r_2] \\
H_{ri} &= A_1\,(\beta/\omega\mu)\tan\phi \left[\frac{a_1 - (b_1/b_2)a_2}{a_2 - (b_2/b_4)a_4}\right][r_2 - (b_2/b_4)r_4] \\
H_{\phi i} &= jA_1\,(\omega\epsilon/\gamma)\,[r_2 + (b_2/b_4)r_4] \\
H_{zi} &= -jA_1\,(\gamma/\omega\mu)\tan\phi \left[\frac{a_1 - (b_1/b_2)a_2}{a_2 - (b_2/b_4)a_4}\right][r_1 + (b_2/b_4)r_4] \\
E_{ro} &= -jA_1\,(\beta/\gamma)\,[(a_1/a_2) - (b_1/b_2)]\,r_4 \\
E_{\phi o} &= -A_1 \tan\phi\,[a_1 - (b_1/b_2)a_2]\,(r_4/a_4) \\
E_{zo} &= A_1\,[(a_1/a_2) - (b_1/b_2)]\,r_2 \\
H_{ro} &= A_1\,(\beta/\omega\mu)\tan\phi\,[a_1 - (b_1/b_2)\,a_2]\,(r_4/a_4) \\
H_{\phi o} &= -jA_1\,(\omega\epsilon/\gamma)\,[(a_1/a_2) - (b_1/b_2)]\,r_4 \\
H_{zo} &= jA_1\,(\gamma/\omega\mu)\tan\phi\,[a_1 - (b_1/b_2)\,a_2]\,(r_2/a_4)
\end{aligned}\right\} \quad \text{(A1)}
$$

The power transported by the slow wave in the z-direction for this helix system is P where

$$
\begin{aligned}
P &= \pi \int_{r=b}^{\infty} (\mathbf{E}\times\mathbf{H}^*)\,r\,dr = \pi \int_{r=b}^{\infty} (E_r H_\phi^* - E_\phi H_r^*)\,r\,dr \\
&= \pi A_1 A_1^*\,(\beta\omega\epsilon/\gamma^2)\left[\int_{r=b}^{a} \{r_2 + (b_1/b_2)r_4\}^2\,r\,dr\right. \\
&\quad + \left\{\frac{(a_1/a_2) - (b_1/b_2)}{(a_2/a_4) - (b_2/b_4)}\right\} \int_{r=b}^{a} \{r_2 - (b_2/b_4)r_4\}^2 r\,dr \\
&\quad + \{(a_1/a_2) - (b_1/b_2)\}\,\{(a_1/a_2) - (b_1/b_4) \\
&\quad \left. + (a_2/a_4) - (b_2/b_4)\} \int_{r=a}^{\infty} (r_4)^2 r\,dr\right] \qquad .. \quad .. \quad .. \quad \text{(A2)}
\end{aligned}
$$

In Equs (A1) and (A2),

$$
\left.\begin{aligned}
a_1 &= I_0(\gamma a) & a_2 &= I_1(\gamma a) & a_3 &= K_0(\gamma a) & a_4 &= K_1(\gamma a) \\
b_1 &= I_0(\gamma b) & b_2 &= I_1(\gamma b) & b_3 &= K_0(\gamma b) & b_4 &= K_1(\gamma b) \\
r_1 &= I_0(\gamma r) & r_2 &= I_1(\gamma r) & r_3 &= K_0(\gamma r) & r_4 &= K_1(\gamma r)
\end{aligned}\right\} \quad \text{(A3)}
$$

ϵ and μ are the permittivity and permeability respectively of free space.

NEW METRE STANDARD FOR N.P.L.

A new instrument for measuring lengths of up to one metre has recently been delivered to the National Physical Laboratory at Teddington.

Developed and produced at the Carl Zeiss Foundation in Western Germany, this instrument measures length standards with an accuracy of 1 part in 10^7 or better. It is known as the Metre Interference Comparator and is one of the only two in existence.

The Comparator uses the wavelength of monochromatic light as the invariable standard of length and the phenomenon of wave-interference as a means of measurement.

BRUSH TO HANDLE SHOCKLEY DIODE

As a result of the recent acquisition of Shockley Transistor by the Clevite Corporation, U.S.A., the Brush Crystal Co. of Hythe, Southampton, have been appointed exclusively to handle the Shockley four-layer diode in the U.K.

The diode is a silicon two-terminal device with two stable states—an 'open' or high-resistance state of more than 1 MΩ and a 'closed' or low-resistance state of a few ohms. It is switched from one state to the other by controlling the voltage across it and the current passing through it.

I.E.E. PART III EXAMINATIONS COURSE

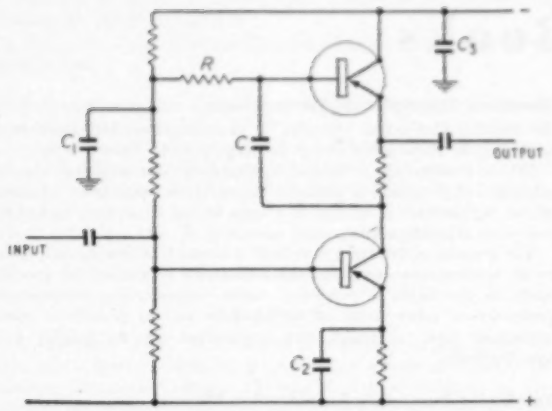
The Department of Electrical Engineering and Applied Physics of the South East London Technical College (Lewisham Way, S.E.4) will be starting a full-time six-months' course for students wishing to sit for Part III Institution of Electrical Engineers' Examinations on 16th October 1961.

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

The Bootstrap Amplifier

SIR,—Following on the original article by W. Tusting in the January issue and a letter from A. Womersley published in April, I would like to draw attention to a transistor circuit which I first used about a year ago and which appears to be a direct equivalent



of Womersley's arrangement. A study of the White cathode-follower suggested it to me.

The circuit is characterized by high gain and low output impedance. Provided R is high in value a relatively small coupling capacitor C can be used. Decoupling capacitors C_1, C_2 and C_3 are used normally.

F. BUTLER.

Prestbury,
Cheltenham,
Glos.
14th April 1961.

Transistor Frequency Response: Graphical Presentation of f_T

SIR,—In their article in your issue of March 1961, Messrs. J. R. James and D. J. Bradley, in dealing with the transistor frequency parameter f_T, have made a number of statements which, it appears to me, could lead to some confusion.

At the beginning of their article they state that f_T is "the frequency at which the small-signal short-circuit current gain in the common-emitter configuration is unity", while it is apparent from their Fig. 1 that the frequency so defined is subsequently referred to as f_1. They cover this discrepancy by stating later in the article that f_1 differs very little from f_T. Although the latter statement may be true for certain types of transistor over a limited range of bias current, it is by no means true universally. At high current levels in particular, the ratio f_1/f_T can be considerable, as was discussed in my article in your December 1960 issue. I consider that the definition given above should be reserved only for the frequency f_1, and that f_T, sometimes called 'transition frequency', is more correctly defined as "the product of the modulus of the common-emitter small-signal short-circuit forward current gain and the frequency at which it is measured, this frequency being so chosen that the modulus is significantly less than its low-frequency value". The latter definition is quite meaningful irrespective of the law of variation of current gain with frequency.

In addition, Messrs. James and Bradley claim that the type of characteristics shown in their Fig. 3 represent a much better switch than those in Fig. 2. This seems to me to be extremely doubtful, since for the load line marked in Fig. 3 the frequency parameter f_T varies over the range 25–500 Mc/s, whereas for the corresponding case for Fig. 2 this range would be approximately 50–350 Mc/s. Apart from the detailed shape of the edge of the pulse these two sets of characteristics are not likely to give very different switching performance over the active region. The authors state further that an approximate value for f_T during the turn-on cycle can be computed graphically from Fig. 3. As they give no further details, I would be interested to know whether they could describe a simple method of doing this. It is possible that the type of characteristics shown in Fig. 3 would show a marked improvement if they were extended to a much higher range of currents.

The user would, of course, be helped by having a set of f_T (V, I) curves over an adequate current range but, if such full information cannot be made universally available, a knowledge of the collector time-constant parameter is probably more useful as a specification for switching transistors.

J. N. BARRY.

The General Electric Company Limited,
Telecommunications and Engineering Laboratories,
Hirst Research Centre,
Wembley, England.
2nd May 1961.

Transistor Circuit Diagrams

SIR,—Mr. T. G. Clark's letter and your own Editorial in the May 1961 issue on Transistor Circuit Diagrams do not, I believe, study the problem deeply enough. As a member of the same Company I have had many discussions with Mr. Clark in the past on this subject and, in common with the rest of the electronics industry, we are not yet agreed!

The basic principles to settle are whether we require a standard direction for the movement of waveforms as has been established on valve circuits (i.e., a positive gradient up the page) or do we prefer our circuits drawn in their most familiar valve form, or is it possible to have both?

With the advent of n-p-n devices of adequate performance and price we are faced, for the first time, with the wholesale use of mixed circuits; in the electronics side of the art it is already commonplace and the trend is soon likely to become universal. The mental agility that Mr. Clark so rightly states is necessary will be required whichever way up the circuit is drawn. The case, therefore, for variation of the complete circuit topography has been overtaken by events.

The case for a 'positive-up' potential gradient is that I consider a consistent presentation of waveforms, with the same convention as valve circuits, as of absolutely fundamental importance. I believe a relatively unskilled engineer must be able to correlate circuit diagram, system block diagram and c.r.o. waveforms with the least possible confusion; waveforms therefore must always be drawn the same way (obviously positive-up), and the circuit must clearly correspond to the waveform. I do not expect senior laboratory engineers to have much difficulty with any system and a small negative sign on an inverted waveform would no doubt be adequate for them, as Mr. Clark asserts, but we cannot give this to a man who is already apt to be confused by polarities, where even a change of oscilloscope is a catastrophe.

However there is a possible compromise to the two extremes; a proposal has been made whereby every stage in a circuit is drawn 'valve way up' so that individual indication of positive or negative h.t. appears at the top as required, but the positive gradient for waveforms is retained (even in complete p-n-p circuits). Some indication is required when the h.t. is in the reverse position and one suggestion is to turn the line and show an arrow pointing in the other direction. This system meets the two main requirements stated above but has serious disadvantages; one is that in parts of the circuit the waveforms appear to run counter to the direction of d.c. potential, another is that apparently similar stages work quite

differently since valve and n-p-n stages conduct more when driven positively at grid or base but p-n-p circuits conduct less. I repeat, p-n-p circuits *are* different, the second fundamental point. To tell a moderately experienced man these are like valve circuits is just postponing the evil day, sooner or later he will have to find out how they work and drawing them the other way up from similar n-p-n circuits clearly indicates the difference in the right manner. Recognizing the circuit upside down is much the lesser of the two evils. I believe the compromise system falls down here but I prefer it to 'variable topography'.

For instruction work of an elementary nature, when converting valve engineers to simple unidirectional transistor circuits, I suggest you publish n-p-n circuits and take advantage of the valve similarity when it is most useful to do so. Postpone the evil day at this stage by all means.

To sum up I believe there are two fundamental points:

(a) To fix waveform polarities is essential, and

(b) p-n-p circuits are different and should be drawn so.

E. W. Kirk.

Decca Radar Ltd.,
Surbiton,
Surrey.
29th May 1961.

New Books

Printed Circuits: Their Design and Application

By J. M. C. Dukes, M.A., D.I.C., A.M.I.E.E. Pp. 228 + xii. Macdonald & Co. (Publishers) Ltd., 16 Maddox Street, London, W.1. Price 40s.

"As it is impossible in a small book to give a comprehensive account of all aspects of printed circuits, attention is specially directed to circuit and equipment design. Sufficient detail, however, is given for this book to be regarded as a first primer on manufacturing techniques and considerable space is given to any fundamental electrical data which is of permanent value.

An appendix enumerates and briefly describes B.S.I. and similar Specifications".

An Introduction to the Theory and Practice of Transistors

By J. R. Tillman, D.Sc., A.R.C.S., and F. F. Roberts, B.Sc.(Eng.), A.M.I.E.E. Pp. 340 + xxii. Sir Isaac Pitman & Sons Ltd., Parker Street, London, W.C.2. Price 57s. 6d.

"Our aim in adding another book has been to present a wider treatment in one volume, mainly by picking out those parts which seem to us to be of most permanent value to a student of the subject, particularly in helping him to obtain a confident quantitative understanding of all the primary physical effects on which minority-carrier transistors depend for their electrical behaviour and an insight into the potentialities and limitations of the devices as circuit elements. It was not our intention to write a handbook for the designer of either devices or circuits."

The Measurement of the Hall Effect with the Aid of Microwaves in Germanium Specimens Changing from N-type to P-type with Changing Temperature

By Tor Stubb. Pp. 19. Published by the Acta Polytechnica Scandinavica Publishing Office, Box 5073, Stockholm 5, Price Sw. Kr. 7.00.

***Stray Losses in Laminated Cores with Particular Reference* to F.H.P. Motors: A Resumé of Published Information**

By W. S. Leung, B.Sc.(Eng.), Graduate I.E.E. Pp. 16. Technical Report Z/T120. Electrical Research Association, Cleeve Road, Leatherhead, Surrey. Price 10s. 6d. (Postage 6d.).

Théorie et Pratique des Circuits de L'Electronique et des Amplificateurs (Vol. 2)

By J. Quinet. Pp. 355 + xv. Published, in French, by Dunod-Editeur, 92 Rue Bonaparte, Paris (6e), France. Price 29 NF.

The Radio Amateur's Handbook 1961 (38th Edn)

By the Headquarters' staff of the American Radio Relay League. Pp. 724. Published by the American Radio Relay League, West Hartford 7, Connecticut, U.S.A. Price $3·50 in U.S.A. ($4·50 elsewhere).

". . . wide dependence on the Handbook is founded on its practical utility, its treatment of radio communication problems in terms of how-to-do-it rather than by abstract discussion.

Virtually continuous modification is a feature of the Handbook—always with the objective of presenting the soundest and best aspects of current practice rather than the merely new and novel."

Electrical Principles of Electronics

By Angelo C. Gillie. Pp. 532 + xix. McGraw-Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 54s. 6d.

"It is written for technical institutes, junior colleges, service schools, and programs in industry that provide training for students whose background is limited to a high school education, including two years of mathematics. . . .

The purpose of this book is to build a broad foundation, adaptable to all applications—to avoid the limitations prescribed by specific study in the fields of servicing, radar, telemetering, microwave, computer or other areas of applications and to provide a comprehensive base on which any application can be quickly and soundly built."

The Radio Amateur's Examination Manual.

Compiled by B. W. F. Mainprise, B.Sc.(Eng.). Pp. 56. The Radio Society of Great Britain, 28 Little Russell Street, London, W.C.1. Price 5s. (By post 5s. 6d.)

"This new R.S.G.C. publication is intended to give guidance to those who aspire to obtain an Amateur (Sound) Licence. The subject matter is treated mainly in question and answer form . . .".

Elsevier's Dictionary of Electronics and Waveguides: Russian Supplement

By W. E. Clason. Pp. 57. Published by Elsevier Publishing Co. and distributed in the U.K. by D. Van Nostrand Co. Ltd., 358 Kensington High Street, London, W.14. Price 42s.

Investigations of Radio Reflections from Satellite-Produced *Ion Trails using 100 Mc CW Radar*

By Martti Tiuri. Pp. 47. Published by Acta Polytechnica Scandinavica Publishing Office, Box 5073, Stockholm 5. Price Sw. Kr. 7.00.

"An investigation has been carried out in which certain echoes observed with a 100 Mc CW radar have been shown to be produced by the ion trail of Sputnik III . . . The observations show that the reflection from the satellite-produced ion trail on 100 Mc is specular. On the basis of the radar cross section found and the duration of the echoes, some estimates of the ion density and shape of the ion trail have been made."

Computer Logic: The Functional Design of Digital Computers

By Ivan Flores. Pp. 458 + xii. Prentice-Hall International Inc., 28 Welbeck Street, London, W.1. Price 45s.

"This book considers computers first at the overall or highest level of organization and proceeds later to the lower levels of detail. We emphasize the organization and functional interrelationship of the fundamental units. These relationships are discussed from the viewpoint of necessity—they are not dependent upon mathematical logic or Boolean Algebra, and they are not arrived at by circuit theory. They are, instead, discussed from the point of view of operational necessity.

A minimum is required of the reader in the way of engineering and mathematical background. The work should be comprehensible to those of some scientific training. Problems have been included to provide a means of acquiring facility with computer theory."

MANUFACTURERS' LITERATURE

Metallurgical and Petrographic Abrasives. Two leaflets published by Geoscience Instruments, U.S.A., give details of their various abrasive papers, cloths and powders and have attached six 1-in. diameter discs coated with silicon carbide of various grit sizes. Details include types of abrasive, disc, strip, sheet and roll sizes, grit sizes and prices.

Geoscience Instruments Corporation, 142 Maiden Lane, New York 38, N.Y., U.S.A.

Plessey Finger Knobs and Mountings. The finger knobs described in this 8-page brochure have been developed in conjunction with Service Research Establishments and have been designed to operate successfully under the most arduous conditions of temperature, humidity, mechanical shock and vibration. In this brochure, types A, S and U are described; these knobs are secured by means of an internally housed collet and avoid the necessity of a transverse hole in the spindle.

The Plessey Co. Ltd., Industrial/Electronic Component Sales, New Lane, Havant, Hants.

RMB Handbook and Catalogue '61. This 84-page handbook gives technical details of the RMB range of miniature, instrument, roller, and special bearings produced by Roulements Miniatures, S.A., of Switzerland. Details included are dimensions, mounting, operation conditions, lubrication and operating life. Distributed in the U.K. by

Miniature Bearings Ltd., 38 Parliament Street, London, S.W.1.

OBITUARY

Mr. Geoffrey Parr, M.I.E.E., died suddenly at his home on 30th May. Born in 1899, he trained as an electrical engineer at Finsbury Technical College. He was Technical Assistant at the Admiralty in 1918-1919 and then became lecturer and demonstrator at The City & Guilds Technical College. In 1926 he joined the Edison Swan Electric Co. and, in 1941, became Editor of *Electronic Engineering*. From 1949 until his death he was Technical Director of Chapman & Hall Ltd.

Mr. Parr was well known for his activities in The Television Society of which he became a Fellow in 1934. From 1936 onwards he was Lecture Secretary and he took on the editorship of the Society's Journal in 1944. He became Hon. Secretary in 1947 and only retired from this a fortnight before his death.

MICROWAVE TRANSISTOR AND AMPLIFIER

A new germanium mesa transistor which operates in the gigacycle-frequency region (10^9 c/s) and as a very wide-band amplifier have recently been developed by Bell Telephone Laboratories.

The device is a diffused-base, alloyed-emitter, p-n-p mesa transistor designed for application as an oscillator at 3 Gc/s or as an amplifier at 1 Gc/s and below. The mesa is 1·8 mils (0·001 in.) long and 1·5 mils wide. Three metal strips, each $\frac{3}{10}$ mils wide by $1\frac{1}{2}$ mils long, are evaporated on to the surface of this tiny plateau and alloyed into the semiconductor. The diffused base of the transistor is only $\frac{3}{50}$ mil thick.

Since conventional encapsulating methods for transistors would introduce too much capacitance and inductance, it is mounted in a coaxial shell which matches a 50-Ω coaxial line. (The germanium wafer is gold bonded to the inner conductor of the output section; the emitter stripe is connected to an internal shield integral with the encapsulation shell; the base stripes are connected to the centre conductor of the input line).

Three-stage amplifiers have been constructed using these transistors and transmission-line principles; they have a gain of 18 dB, flat within 1 dB over the frequency band from below 1 Mc/s to over 750 Mc/s.

In the amplifier, the emitter is earthed in order to maintain outer-conductor continuity, and the transistor biases are stabilized with d.c. local-shunt feedback. A low-frequency transistor is inserted in each biasing circuit to increase the feedback at d.c. without wasting appreciable amounts of high-frequency gain.

The amplifiers have shown excellent stability and the noise figure measured at 200 Mc/s is 5·5 dB with the feedback loops open.

NETWORK THEORY SYMPOSIUM

A residential symposium on Network Theory is to be held by the Department of Electrical and Control Engineering of The College of Aeronautics, Cranfield, from 18th to 22nd September 1961. Further details and registration forms are obtainable from S. R. Deards, Department of Electrical and Control Engineering, The College of Aeronautics, Cranfield, Bletchley, Bucks.

BIRTHDAY HONOURS

Knight Bachelor

Allen George Clark (Chairman and Managing Director, Plessey Co. Ltd.).

C.B.

Air Vice-Marshal T. U. C. Shirley (Deputy Controller, Electronics, Ministry of Aviation).

A. Wolstencroft (Director of Radio Services, G.P.O.).

C.B.E.

A. F. Lampitt (Managing Director, Telephone Cables Ltd.).

W. H. Penley (Deputy Chief Scientific Officer, Royal Radar Establishment).

N. H. Searby (Manager, Guided Weapons Department, Ferranti Ltd.).

F. C. Williams (Professor of Electrical Engineering, University of Manchester).

O.B.E.

E. A. Hamilton-Hill (Director and General Manager, Rediffusion [Malta]).

M.B.E.

W. E. Bell (Superintendent, Applied Electronics Laboratory, G.E.C. Portsmouth).

R. G. Hawkins (Chairman and Managing Director, Redcliffe Radio & Engineering Co. Ltd.).

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for May 1961*

Date 1961 May	MSF 60 kc/s 1500 U.T. Parts in 10^9	Droitwich 200 kc/s 1030 U.T. Parts in 10^9
1	— 15·0	+ 12
2	— 15·5	+ 12
3	— 15·3	+ 12
4	— 14·7	+ 14
5	— 14·8	+ 12
6	— 14·6	+ 13
7	— 14·2	+ 13
8	— 14·3	+ 14
9	— 14·7	+ 15
10	— 14·9	+ 15
11	— 15·0	+ 15
12	NM	+ 15
13	— 15·7	+ 14
14	— 15·2	+ 13
15	— 15·5	+ 14
16	— 15·0	+ 14
17	— 15·2	+ 13
18	— 14·7	+ 13
19	— 15·2	+ 12
20	— 15·1	+ 13
21	— 15·2	+ 13
22	— 15·0	+ 12
23	— 15·2	+ 14
24	— 15·2	+ 14
25	— 15·1	+ 14
26	— 15·1	+ 15
27	— 15·2	+ 14
28	— 15·3	+ 15
29	— 15·4	+ 14
30	— 15·3	+ 14
31	— 15·2	+ 15

* *Nominal frequency corresponds to a value of* 9 192 631 770 *c/s for the caesium F,m (4,0)—F,m(3,0) transition at zero field. NM = Not measured.*

New Products

Miniature Silicon Diode

A series of silicon diodes, which can be used as general-purpose diodes or low-power rectifiers, has been developed by the Electronic Apparatus Division of A.E.I.

The basic diode is of double-ended construction and is enclosed in a hermetically-sealed envelope measuring $\frac{1}{10}$ in. diameter by $\frac{1}{4}$ in. long.

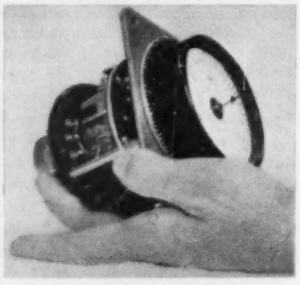
The series, known as types MS1H to MS5H, covers p.i.v. ratings from 60 to 300 V. All types are rated at 250 mA maximum forward current and, at 25 °C with a forward current of 200 mA, the maximum potential drop is 1 V. At the diode's maximum permissible junction temperature of 150 °C and rated p.i.v., the maximum reverse leakage current is 50 μA for the 60-V and 100-V diodes and 100 μA for the 150-V, 200-V and 300-V diodes; at 25 °C, the maximum leakage is 1 μA for all types. The maximum junction capacitance is 5 pF and the stored charge during inverse recovery is 20 mμC.

Associated Electrical Industries Ltd.,
Electronic Apparatus Division,
Carholme Road, Lincoln.

Automatic Synchronous Timer

An automatic timer which is designed to meet the majority of common industrial control applications requiring single-circuit timing is the latest addition to the Elremco range of timers.

Two basic models are available, the Star and the Star/D; the Star is arranged to

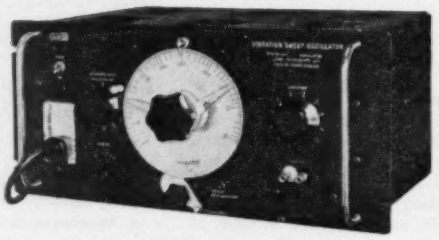


provide a s.p.d.t. switch changeover for a pre-set time (during the timing period) and the Star/D is arranged to provide a s.p.d.t. switch changeover after a pre-set time (at the end of the timing period). In both models the switching capacity is 5 A at 240 V a.c. to $\frac{1}{2}$ A at 440 V a.c. Full-scale time ranges are available from 0—12 sec to 0—24 hr., and models can be supplied for operation from the normal 50 to 60-c/s mains supplies (110 V—440 V).

Electrical Remote Control Co. Ltd.,
Bush Fair,
Harlow, Essex.

Automatic L.F. Sweep Generator

A resistance-capacitance tuned sweep oscillator covering in a single-frequency



range 5 c/s to 5 kc/s has been announced by Dawe Instruments.

The instrument has been designed to drive a power amplifier and a moving-coil generator for vibration resonance search and endurance tests as required in component and equipment specifications, such as British Standards B.S. 2011 : 1954 and G 100, Inter-Service Specifications R.C.S.11, DEF-5011 and K.114; as well as the International I.E.C. Publications No. 68.

In the oscillator, the frequency sweep control is driven by a motor which is controlled to provide six sweep rates from 0·2 to 10 octaves/min. The sweep can be set to cover the whole of the frequency range or any selected part of the range.

A push-pull output is provided at three output terminals; each output being continuously variable up to 12 V r.m.s. into a load resistance of 10 Ω. (The photograph shows a pre-production model with only two output terminals.)

Dawe Instruments Ltd.,
Harlequin Avenue,
Great West Road,
Brentford, Middx.

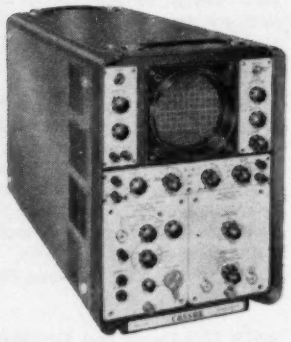
High-Speed Multi-Purpose Oscilloscope

The latest addition to the Cossor range of oscilloscopes is Model 1076, which is designed as a basic single-beam instrument with a number of plug-in units for both the X and Y channels to extend the performance.

The basic units of the oscilloscope are a single-beam c.r.t., Y-deflection amplifier, X-deflection amplifier, time-base generator, which has calibrated speeds over a range of $2{\cdot}5 \times 10^8$: 1, and power supplies. The 5-in. precision c.r.t., operating at 10 kV, provides a linear display area of 6 cm × 10 cm.

The performance of the oscilloscope, when the wide-band preamplifier model 1078 is used, is as follows: Y amplifier: bandwidth, d.c. to 60 Mc/s (30% down), 2 c/s to 60 Mc/s at a.c. input; rise-time 6 mμsec to 7 mμsec; sensitivity 50 mV/cm to 20 V/cm calibrated; input impedance 1 MΩ/20 pF on all ranges, 50 Ω or 75 Ω termination by switch selection. Signal delay is 150 mμsec.

The time-base has 24 calibrated ranges from 0·02 μsec/cm, and sweep expansion of × 1 or × 5 on all ranges from 0·1 μsec/cm to 10 sec/cm. A d.c.-coupled brightening pulse to the c.r.t. grid gives uniform bright-



ness along the trace length and complete flyback suppression on all ranges.

With the trigger unit, model 1079, the X-amplifier bandwidth is d.c. to 2 Mc/s (30% down) and the sensitivity is 1 V/cm to 100 V/cm (calibrated). The trigger source selection can be set for positive or negative signals from the Y amplifier, external input or 50 c/s. The trigger system permits four modes of operation: h.f. sync, automatic, a.c. and d.c. and threshold and level controls are incorporated.

For calibration, a square-wave source is included to provide eighteen fixed voltage levels from 0·2 V to 100 V peak to peak.

Other plug-in units include a high-gain differential Y-preamplifier and a precision delayed-trigger unit.

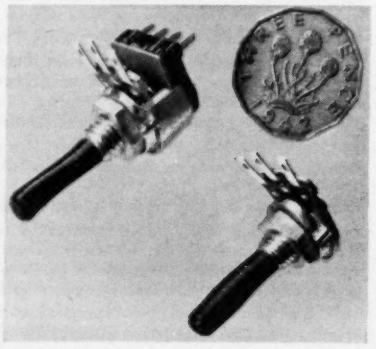
Cossor Instruments Ltd.,
Highbury Grove, London, N.5.

Sub-Miniature Potentiometers

Two sub-miniature potentiometers, the first of a range designed for use in transistor circuits, have been announced by Egen.

Robustly constructed with a die-cast body of $\frac{1}{2}$ in. diameter, these bush-mounted controls have a tough plastics-moulded spindle of $\frac{3}{16}$ in. max. diameter; they will be available with standard resistance values in the range 1 kΩ to 3 MΩ, log or linear law.

The first potentiometer, Type 363, is not fitted with a switch: the second, Type 365, is fitted with a d.p.d.t. switch rated at 50 V



150 mA. Both models are constructed with standard soldering tags.

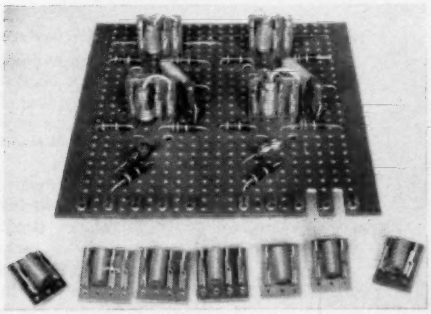
Other models for printed-wiring panels are planned for production in the near future.

Egen Elecric Ltd.,
Charfleet Industrial Estate,
Canvey Island, Essex.

Semiconductor Holders

Two holders which have been designed to take transistors and diodes have recently been announced by Vero Electronics.

These holders are basically rectangles of insulating board on which are mounted contacts and a tubular socket for the semiconductor. The diode version has two contacts and the transistor version three. One end of each contact has a slot and a hole to facilitate the soldering of the semiconductor lead, while the other end is formed into a rectangular cross-section pin



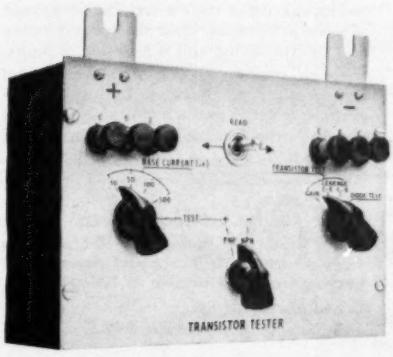
to enable the holder to be plugged into a socket. The holders will mount semiconductors of 5 to 6 mm. diameter.

Vero Electronics,
South Mill Road, Regent's Park, Southampton.

Transistor Tester

Grundy & Partners have produced a tester for n-p-n and p-n-p transistors which has been designed for use in conjunction with all multi-stage meters of 1-mA basic movement or better, preferably with inbuilt overload protection. The mounting terminals are appropriately spaced to enable the unit to be mounted directly on to a Universal Avo Meter.

Transistor measurements include collector–emitter and collector–base leakage current at a potential of 4·5 V. A useful measurement of current gain can be made of transistors up to 800-mW dissipation and a reasonable indication of gain is given for higher-power types.



At a fixed collector potential of 4·5 V, four switched base input current ranges of 10, 50, 100 and 500 μA are provided.

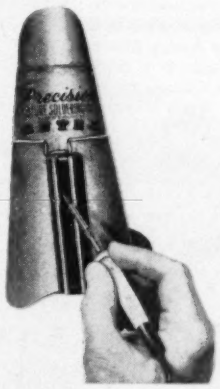
Diodes are tested in the forward direction by passing a current of up to 10 mA through them; reverse current is checked at a potential of 9 V.

Provision is also made for testing the internal batteries under load.

Grundy & Partners Ltd.,
3 The Causeway, Teddington, Middx.

Soldering-Iron Bit Cleaner

A bit cleaner for attachment to the standard A.N.T.E.X. soldering-iron rest has recently been introduced.



This is a steel-wire brush in a mounting which can easily be fitted to the stand by two bolts; the bolt holes in the stand are those also used for the supporting bracket.

The brush (approx. 3½ in. × ½ in.) is held in its mount by spring clips and can therefore be replaced as a separate item.

A.N.T.E.X. Ltd.,
7-8 Idol Lane, London, E.C.3.

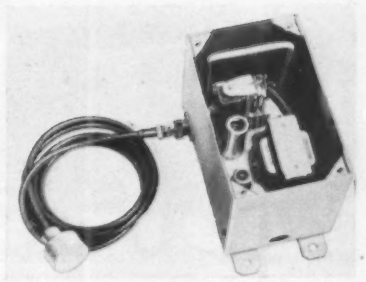
Inductive Proximity Switch

A proximity switch which has been designed as an alternative to a mechanical limit switch has been introduced by Donovan Electrical.

It is the type CE32 and it comprises an inductive sensing unit of approx. 1½ in. max. diam. by 1½ in., connected to a control unit by a 6-ft. flexible lead. The switch is capable of up to five operations per second when the internal relay is used.

In the proximity switch, a t.a.t.g. oscillator circuit operates at 5 Mc/s with the sensing-head inductor incorporated in the oscillator feedback circuit. When metal is in the vicinity of the sensing head, losses are coupled into the feedback circuit and oscillation ceases. This causes a rise in the anode current of the oscillator valve and the operation of a relay. The relay is normally fitted with two changeover contacts rated at 250 V, 5 A. Operating from either 230 or 400 V a.c., the control unit measures 7⅜ in. × 5⅞ in. × 4 9/16 in.

The proximity switch will detect metal objects up to approximately ⅜ in. away from the sensing head and, when the object being detected is roughly near the same area as the



face of the sensing head, the repetitive accuracy is of the order of 0·02 in.

The Donovan Electrical Co. Ltd.,
Safuse Works, Northcote Road,
Stechford, Birmingham 33.

Transistorized Portable Tape Recorder

A transistorized portable tape recorder has been announced by E.M.I. Electronics.

Known as the RE321, it is a single-track machine designed to operate at a tape speed of 7½ i.p.s. The frequency response on record is substantially flat from 50 c/s to 10 kc/s and the input sensitivity is 75 μV for peak recording level; the input impedance is 30 Ω.

This instrument is completely self-contained and operates from eight U2-type batteries which enables the recorder to be

used intermittently for about two hours before battery replacement is necessary. An associate company of E.M.I., Alpha Batteries, are shortly to produce a Rayovac U2 battery which has approximately three times the life of a conventional U2.

The RE321 measures 8 in. × 14 in. × 6¾ in. and weighs approximately 17 lb.

E.M.I. Electronics Ltd.,
Hayes, Middx.

A.C.-Converter Voltmeter

The Solartron a.c.-converter voltmeter, type LM903, measures a.c. voltages (0–500 V) over the frequency range 20 c/s to 5 Mc/s. It was primarily designed to operate in conjunction with the digital voltmeter type LM902.2 to give a four-figure digital display of a.c. voltages from 100 μV to 500 V. The measuring accuracy, when used with the LM902.2, is ±0·25% (50 c/s

to 100 kc/s) and ±2% (20 c/s to 750 kc/s) of full scale on each range, with useful indications up to 5 Mc/s.

Operating alone, the LM903 covers 0–15 mV to 0–500 V in ten ranges, with a ±2% measuring accuracy on the built-in meter. The input impedance is 1 MΩ shunted by approximately 40 pF.

Pre-set controls and an internal calibration voltage permit accurate initial setting of the instrument and the amplified a.c. signal is connected to a coaxial socket on the front panel so that it can be monitored.

Although supplied as bench units, the a.c. converter LM903 and the digital voltmeter LM902.2 are easily adapted for standard rack-mounting. Both instruments operate from 110/240 V, 50–60-c/s mains. The dimensions are 9½ in. × 18 in. × 16 in. (bench models) and 8¾ in. × 19 in. × 16 in. (rack-mounting models); the weight is 43 lb.

Solartron Laboratory Instruments Ltd.,
Cox Lane, Chessington, Surrey.

Ultra-Violet Photomultiplier

Mullard have introduced a modified version of the photomultiplier tube type 56AVP in which the spectral response has been extended well into the ultra-violet region.

The new tube, type 56 UVP, is fitted with a quartz end-window for use with gas scintillation counters. It is similar to the 56 AVP but its spectral response extends into the u.v. region and it has a useful response down to 2,300 Å, with a peak at 4,000 Å (0·4 μ).

For high-definition work, very small electron transit-time differences are essential and in the 56 UVP they have been kept down to 2 mμsec at half-pulse height, with a rise-time of the same duration.

This reduction of transit-time differences is due to the input electron-optical system in which the photocathode is curved so that the path lengths to the first multiplier stage are approximately equal for electrons leaving any part of its useful area. The transit-time difference between electrons arriving at the first dynode from the periphery of the photocathode and those arriving from the centre is 0·3 mμsec.

Uniformity of transit times through the 14 multiplier stages is maintained by careful shaping of the dynodes and by the additional focusing electrodes situated between each multiplier stage.

Mullard Ltd.,
Torrington Place, London, W.C.1.

Heavy-Duty Transistorized Power Supply

A recent addition to the Gossen range of power supplies has been introduced into the U.K. by Aveley Electric.

The unit is a fully-transistorized, high-stability d.c. source for operation from a 40

to 60-c/s input supply. It has continuously-variable voltage range from 5 V d.c. (12 A) to 60 V d.c. (5·4 A). The change from no-load to full-load can be made in 0·2 msec and the ripple potential is 1 mV.

The power supply incorporates an electronically-controlled automatic circuit-breaker, thermal-overload protection and meters to monitor the output voltage and current. Maximum permissible ambient temperature at full load is 30 °C.

Distributed in the U.K. by

Aveley Electric Ltd.,
Ayron Road,
Aveley Industrial Estate,
South Ockendon, Essex.

Component Identification Printer

An automatic printing machine which has been designed primarily for printing identification marks on to small-diameter cylindrical components with coaxial-wire leads has been introduced by Rejafix.

The machine will handle components (with coaxial-wire leads) of ½ in. to 3⅜ in. body length and 0·08 in. to ⅞ in. body diameter with a maximum length of wire protruding from each end of 2⅞ in.

Components can be marked with letter and/or figure coding or they can be colour-banded. The print can be placed anywhere along the length of the body of the articles and any printing area up to the maximum of 3 in. × 2⅜ in. is available.

The output of the machine is up to 3,600 components per hour, depending upon the operator's skill in feeding, except when used for colour-banding, when an output of up to 3,000 per hour can be expected.

When modified for handling articles without wires, the machine will take objects from 1/16 in. to ½ in. diameter and up to the maximum length of 10 in., with a maximum length of print of 3 in.

Rejafix Ltd.,
81–83 Fulham High Street,
London, S.W.6.

Industrial Pen and Hot-Stylus Recorders

A range of industrial direct-writing recorders, manufactured by Officine Galileo of Florence, for the permanent recording of a wide range of l.f. phenomena, is now available in the U.K. through Leland Instruments.

These direct-writing instruments, employing ink pen or hot stylus, can be supplied in several standard models—portable single pen, two- or four-pen recorders, or rack-mounted on wheels and having 2, 4, 6, 8, 16 or 32 channels, thus permitting the simultaneous recording of an equivalent number of phenomena.

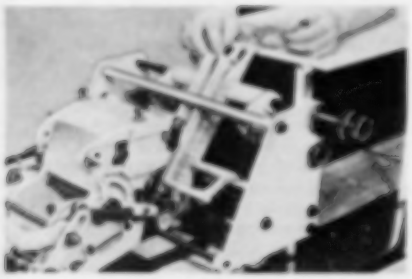
The interchangeable pre-amplifier units and (if required) filters incorporate printed

circuits and transistors. A range of pre-amplifiers is available to cover most applications.

The hot-stylus record has the advantage of being in Cartesian co-ordinates permitting easy interpretation and correlation of multi-channel recording. The multi-channel instruments can have a built-in photographic recorder and a mirror galvanometer allowing the examination of phenomena whose higher-frequency components (200–800 c/s) would not be reproduced by direct-writing pens.

The Galileo recorders operate from a power supply of 110 to 220 V 50 c/s, have a response time of 0·01 sec and paper speeds varying from 2·5 mm/sec to 100 mm/sec.

Leland Instruments Ltd.,
145 Grosvenor Road, London, S.W.1.



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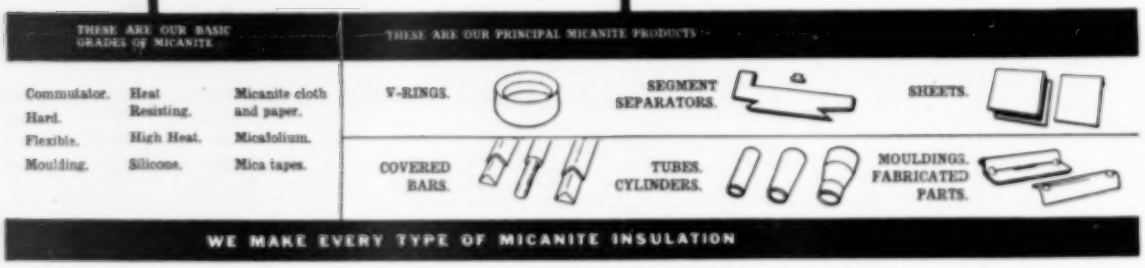

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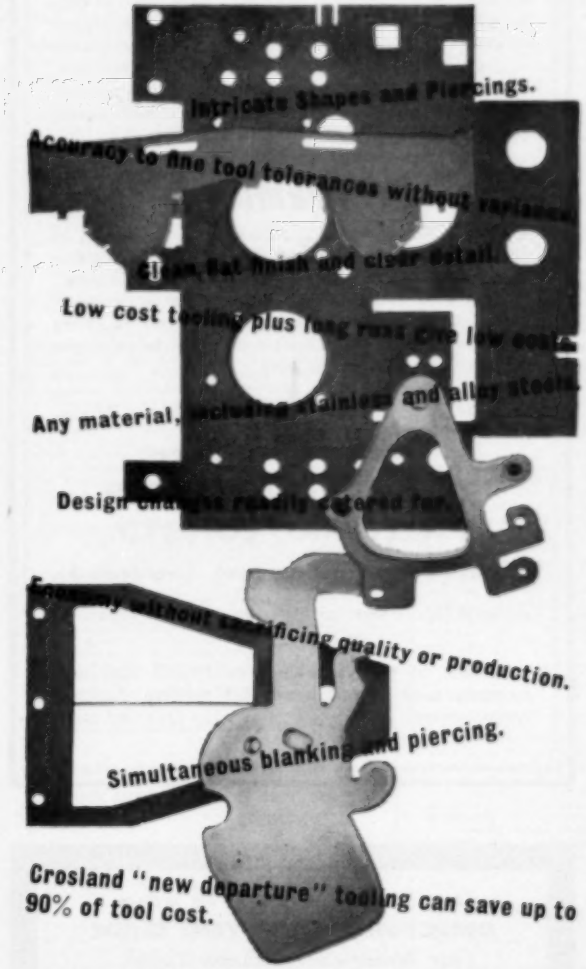
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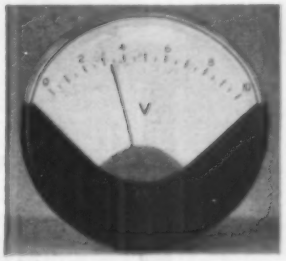

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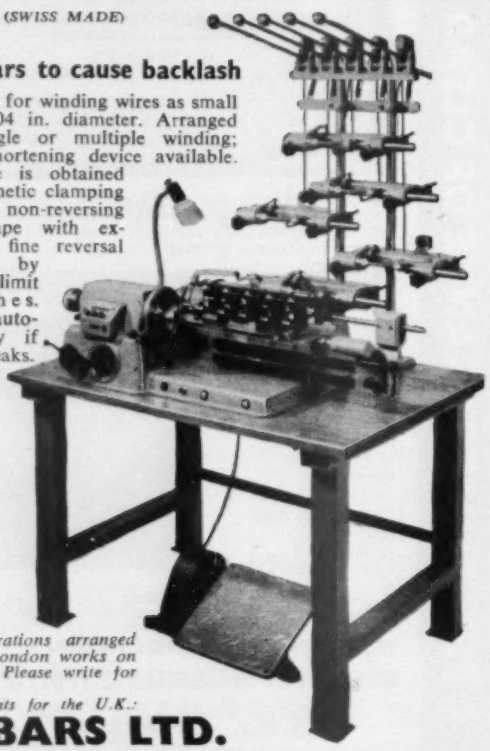

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NEW TV TUNER HIGH SLOPE VHF TETRODE

EDISWAN MAZDA 30F27

The 30F27 is a frame grid VHF tetrode having a mutual conductance of 15 mA/V at an anode current of 13·5 mA and a screen current of 1·7 mA with variable-mu characteristics to reduce cross-modulation effects at high signal levels.

This tetrode used in the RF stage of a television tuner offers certain advantages over the more conventional double triode cascode arrangement. For instance, the number of circuit components required is smaller, the layout is simpler and for valves having comparable slopes the single cathode type valve can be manufactured more economically.

Normally the noise performance of a tetrode is inferior to that of a triode due to the presence of partition noise arising from the screen current. However the 30F27 has been specially designed to provide a low ratio of screen to anode current to minimise the effect of partition noise while still retaining good screening between control grid and anode. When this is used in conjunction with frame-grid techniques a high slope per milliamp of anode current can be obtained resulting in a high gain RF valve with a noise performance much superior to that of a conventional pentode and equal to that of a double triode cascode amplifier such as the 30L1.

Heater Current (amps)	I_h	0·3
Heater Voltage (volts)	V_h	3·7

TENTATIVE RATINGS AND DATA

Maximum Design Centre Ratings

Anode Dissipation (watts)	$P_{a(max)}$	2·5
Screen Dissipation (watts)	$P_{g2(max)}$	0·4
Anode Voltage (volts)	$V_{a(max)}$	250
Screen Voltage (volts)	$V_{g2(max)}$	230
Heater to Cathode Voltage (volts rms)	$V_{h-k(max)rms}$	90*
Cathode Current (mA)	$I_{k(max)}$	18

* From Cathode to higher potential heater pin.

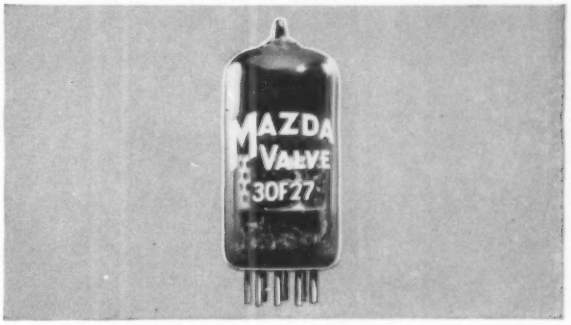
Inter-electrode Capacitances (pF)†

Input Capacitance	C_{in}	6·3
Output Capacitance	C_{out}	1·8
Grid 1 to Anode	C_{g1-a}	0·027
Grid 1 to Grid 2	C_{g1-g2}	2·0
Grid 1 to Cathode	C_{g1-k}	4·0

† Measured in fully shielded socket, without can.

Maximum Dimensions (mm)

Overall Length	56
Seated Height	49
Diameter	22·2



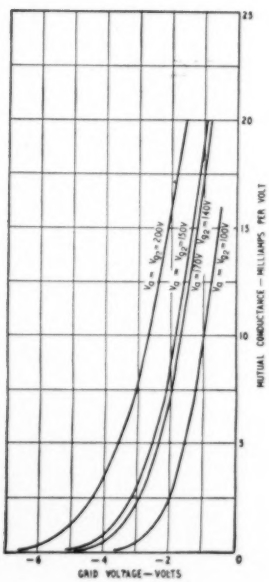
TYPICAL OPERATION

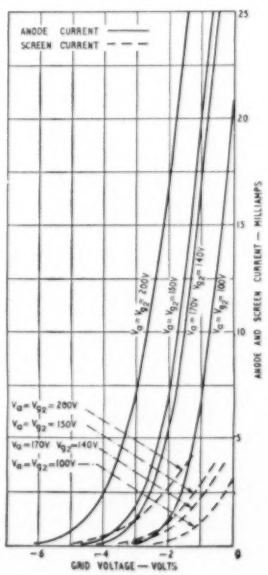
		Cathode Self Bias Circuit	Grid Current Bias Circuit
Supply Voltage (volts)	V_b	200	200
Anode Voltage (volts)	V_a	170	150
Screen Voltage (Initial) (volts)	V_{g2}	140	105
Anode & Screen Common Decoupling Resistor (kΩ)		—	3·3
Anode Decoupling Resistor (kΩ)		2·2	—
Screen Decoupling Resistor (kΩ)		33	33
Cathode Bias Resistor (Ω)	R_k	82	—
Grid Current Bias Resistor (kΩ)	R_{g1}	—	330
Grid Bias Voltage approx. (volts)	V_{g1}	−1·25	—
Anode Current (mA)	I_a	13·5	14
Screen Current (mA)	I_{g2}	1·7	1·4
Mutual Conductance (mA/V)	g_m	15	15·5
Inner Amplification Factor (g_1 to g_2)	μ_{g1-g2}	60	—
Equivalent Grid Noise Resistance (Ω)	R_{eq}	450	—
Input Loss at 50 Mc/s (kΩ)	$r_{gl-k(w)}$	6·8¶	—
Input Capacity Working (pF)	$C_{in(w)}$	10·3¶§	—
Change in Input Capacity produced by biasing valve to cut-off (pF)	$\triangle C_{in(w)}$	2·9¶	—

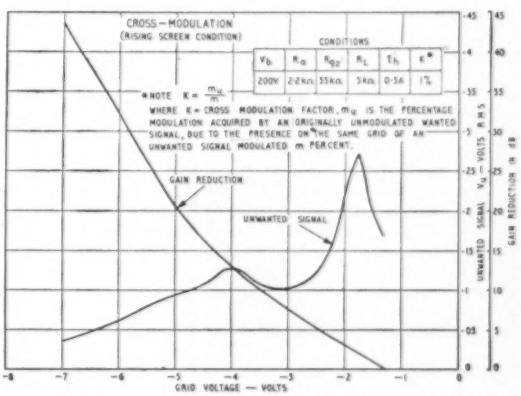
¶ Measured at 50 Mc/s with the three cathodes strapped and taken directly to earth.

§ Inter-electrode capacity with holder capacity balanced out.

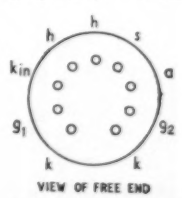
Tentative Characteristic Curves of Ediswan Mazda Valve Type 30F27

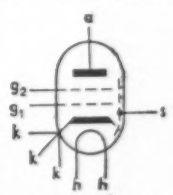






Base: B9A (Noval) **Mounting Position**: Unrestricted





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